NAVPERS 93400A-6

# FUNDAMENTALS OF ELECTRONICS

## VOLUME 6

## MICROWAVE CIRCUIT APPLICATIONS



Bureau of Naval Personnel

# PREFACE

This book is part of a nine-volume set entitled "Fundamentals of Electronics". The nine volumes include:

Volume 1a - NavPers 93400A-1a, Basic Electricity, Direct Current
Volume 1b - NavPers 93400A-1b, Basic Electricity, Alternating Current
Volume 2 - NavPers 93400A-2, Power Supplies and Amplifiers
Volume 3 - NavPers 93400A-3, Transmitter Circuit Applications
Volume 4 - NavPers 93400A-4, Receiver Circuit Applications
Volume 5 - NavPers 93400A-5, Oscilloscope Circuit Applications
Volume 6 - NavPers 93400A-6, Microwave Circuit Applications
Volume 7 - NavPers 93400A-7, Electromagnetic Circuits and Devices
Volume 8 - NavPers 93400A-8, Tables and Master Index

If you are becoming acquainted with electricity or electronics for the first time, study volumes one through seven in their numerical sequence. If you have a background equivalent to the information contained in volumes one and two, you are prepared to study the material contained in any of the remaining volumes. A master index for all volumes is included in volume eight. Volume eight also contains technical and mathematical tables that are useful in the study of the other volumes.

A question (or questions) follows each group of paragraphs. The questions are designed to determine if you understand the immediately preceding information. As you study, write out your answers to each question on a sheet of paper. If you have difficulty in phrasing an answer, restudy the applicable paragraphs. Do not advance to the next block of paragraphs until you are satisfied that you have written a correct answer.

When you have completed study of the text matter and written satisfactory answers to all questions on two facing pages of the book, compare your answers with those at the top of the next even-numbered page. If the answers match, you may continue your study with reasonable assurance that you have understood and can apply the material you have studied. Whenever your answers are incorrect, restudy the applicable material to determine why the book answer is correct and yours is not. If you make an honest effort to follow these instructions, you will have achieved the maximum learning benefits from each study assignment.

Follow the directions of your instructor in answering the review questions included at the end of each chapter.

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# CHAPTER 49

# UHF TRANSCEIVER

The name TRANSCEIVER was derived by combining the words "transmitter" and "receiver." As this name suggests, a transceiver consists of a transmitter and a receiver, arranged and connected so as to function as a single unit.

Although transceivers can be constructed to operate at nearly any radio frequency, they are more popular in the high frequency bands. The transceiver to be described in this chapter operates within the band of frequencies extending from 300 Mc to 3000 Mc.

Due to the short wavelength of the signals at high radio frequencies, much smaller values of circuit inductance and capacitance are required than were used in previously discussed circuits. Because of this, new techniques of circuit and component construction become necessary. It must be stressed, however, that even though the tubes, tank circuits, etc., may have unusual physical appearances, the basic theory of operation remains the same. This chapter is designed to consider some of the problems encountered in a typical UHF transceiver, and how they are solved.

## 49-1. Characteristics of UHF Systems

The frequencies from 300 Mc to 3000 Mc constitute what is called the ULTRA HIGH FREQUENCY (UHF) band. The location of this band of frequencies with respect to the other frequency bands in the radio spectrum is shown in Table 49-1.

| BAND | FREQUENCY (Mc) | ABBR. |
|---|---|---|
| Very low freq. | 0.01 to 0.03 Mc | VLF |
| Low freq. | 0.03 to 0.3 Mc | LF |
| Medium freq. | 0.3 to 3 Mc | MF |
| High freq. | 3 to 30 Mc | HF |
| Very high freq. | 30 to 300 Mc | VHF |
| Ultra high freq. | 300 to 3,000 Mc | UHF |
| Super high freq. | 3,000 to 30,000 Mc | SHF |
| Extremely high freq. | 30,000 to 300,000 Mc | EHF |

Table 49-1 - Radio-Frequency Spectrum.

Notice that the RF spectrum commences at 0.01 megacycles per second (10 kc) and progresses through successively higher bands to the extremely high frequency band which ends at 300,000 Mc.

In the early periods of radio communications development, few stations were in operation and most of the radio spectrum was vacant. As thousands of additional stations were placed on the air (commercial, military, amateur, police, etc.) signals from one transmitter began to interfere with those from another due to crowding of the stations. To minimize this problem, new communications services (such as television, FM, citizens band, etc.) were assigned to higher and less crowded frequency bands as they came into existance. Thus, the frequency bands, as listed in Table 49-1, were developed and assigned.

A radio wave, or electromagnetic wave, which leaves a transmitting antenna may travel to a receiving antenna in more than one way. The transmitted radio wave may be transferred to the receiving antenna by a ground wave (sometimes referred to as a surface wave), by a sky wave, or by a space wave. Although other transfer mechanisms exist, the ones listed above are the most important.

Since the ground waves travel over the surface of the earth they are greatly affected by the conductivity of the earth and any obstructions such as mountains or buildings on its surface. As it passes over the ground, the surface wave induces a voltage in the earth, setting up eddy currents. The energy to establish these currents is absorbed from the surface wave, thereby weakening it as it moves away from the transmitting antenna. Increasing the frequency of the radio wave, rapidly increases the attenuation so that surface wave communication is limited to relatively low frequencies.

That part of the radio wave that moves upward and outward and is not in contact with the ground is called the sky wave. Some of the energy of the sky wave is refracted by the ionosphere so that it comes back towards the earth. A receiver located in the vicinity of the returning sky wave will receive strong signals even though several hundred miles beyond the range of the ground wave. However, as the frequency of the radio wave is increased the ionosphere refracts less energy back to the earth and communication at ultra high frequencies, by use of the sky wave, becomes very unreliable.

A radio wave that travels directly from the transmitting antenna to the receiving antenna, in the atmosphere adjacent to the earths surface (first 10 miles above the surface), is called the space wave. Space wave energy may reach the receiving antenna by direct travel or as the result of reflection or refraction. Radio communications in the UHF range are normally accomplished by space-wave propagation.

In general, the VHF and UHF waves follow approximately straight lines. Large hills or mountains cast a radio shadow over these areas in the same way that they cast a shadow in the presence of light rays. A receiver in a radio shadow will receive a reduced signal, or in some cases no signal at all. Theoretically, the range of contact is the distance to the horizon, and this distance is determined by the heights of the antennas. This is why UHF communication is known as line-of-sight communication.

**The antennas used for UHF communications can be made highly directive. The reason for this is the small size of the antenna elements. More elements can be used in a relatively small space. Because it is line-of-sight and can be highly directive, relay stations are often used to extend the range of communications.**

One of the advantages of using UHF communication is found in the physical size of the unit. The high operating frequency and low power required of the UHF transceiver result in low values of components. The low values permit a reduction in the physical size of individual components and, hence, in the size of the entire unit. This compactness makes the UHF transceiver highly suitable for air-to-air and ship-to-ship communications.

The block diagram of the basic transmitter section of a UHF transceiver is shown in Figure 49-1. Notice the similiarity between the block diagram of the UHF transmitter and the basic transmitter discussed in previous chapters. Waveforms indicating the input and output of each block are shown in Figure 49-1. The oscillator block generates a radio frequency signal which is amplified by the buffer amplifier. The buffer amplifier also isolates the oscillator,

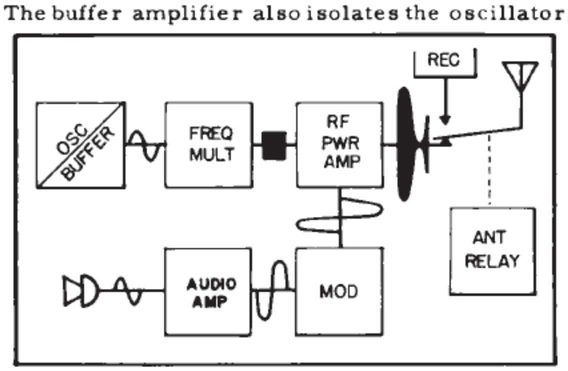

Figure 49-1 - Basic UHF transmitter.

or prevents it from being loaded by the following stages. The constant amplitude radio frequency signal from the buffer amplifier is then applied to a frequency multiplier, where its frequency is increased. The frequency multiplier may be a doubler, tripler, or quadrupler, depending on the desired carrier frequency. The output of the multiplier stage is then applied to the RF power amplifier.

The microphone converts mechanical energy into electrical energy which, in this case, varies at an audio rate. This small audio signal is then applied to audio amplifiers which increase its amplitude and power level. The audio output from the modulator is then applied to one of the elements of the RF power amplifier, the specific element depending on whether low or high level modulation is being used. The output of the RF power amplifier is a modulated RF signal of high power.

Figure 49-1 shows that the transceiver contains an antenna relay which permits connection of a single antenna to either the transmitter or the receiver. In its normal position the antenna relay is deenergized and the antenna is connected to the receiver section. However, when transmitting the antenna relay is energized and the antenna is connected to the transmitter. Figure 49-1 shows the antenna relay in its energized position.

Figure 49-2 shows the block diagram of the receiver section of the UHF transceiver. The antenna relay is shown in its normal (deenergized) position, which connects the antenna to the receiver. The first stage of the UHF receiver is the RF amplifier. This is a tuned amplifier which raises the level of the input signal. After amplification, the signal is then sent to the mixer. The other input to the mixer is a constant frequency, constant amplitude signal. It is produced by a stable local oscillator. The output from the oscillator is applied to a frequency multiplier. As in the transmitter, the multiplier may be a doubler, tripler, or quadrupler. It may consist of several stages of frequency multiplication. At the mixer, heterodyning between the modulated input signal and the constant frequency oscillator signal take place. Of the frequencies produced by this action only one, the intermediate frequency, is chosen and applied to the first IF amplifier. Normal UHF receivers have more than one stage of IF amplification. It is in these stages where the gain and bandwidth of the receiver are determined. After the signal has been sufficiently amplified, it is then applied to the detector stage where demodulation takes place. The signal now is composed of audio frequencies. The output from the detector is then applied to an audio amplifier. The number of audio amplifiers used is determined by the requirements of

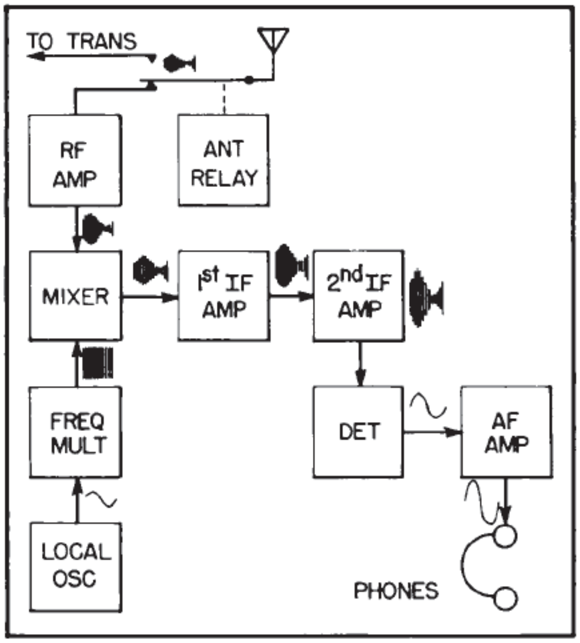


Figure 49-2 - Basic UHF receiver.

the user. In most UHF receivers, more than one stage of audio amplification is used. The output of the audio amplifier is then applied to a reproducer, which may be headphones as shown in the figure or a speaker.

Q1. What is the function of the frequency multiplier in both the receiver and the transmitter?

TRANSCEIVER OSCILLATORS

49-2. Tuned Plate Tuned Grid Oscillator

Due to the high operating frequency the local oscillator of a UHF receiver must have better frequency stability than its broadcast band counterpart. For this reason UHF receivers commonly use a CRYSTAL OSCILLATOR to achieve a high degree of frequency stability. The crystal oscillator is similar to that of the TUNED PLATE TUNED GRID (TPTG) OSCILLATOR. In order to better understand the operation of the crystal oscillator, the operation of the TPTG oscillator will be analyzed first.

The tuned plate tuned grid oscillator as shown in Figure 49-3 is characterized by parallel resonant circuits in both the grid and plate circuits. At first glance, the circuit looks like nothing more than a tuned amplifier, and it is at some frequencies. This qualification is made because a visual examination reveals the absence of a feedback path. Without feedback of the proper phase, the circuit will not oscillate.

Feedback would be provided if a small capacitor were physically connected between the plate and grid of the tube. In that way a portion of the plate signal would be fed back to the grid. However, there is no need to physically connect a capacitor between plate and grid because the triodes grid-to-plate capacitance, $C_{gp}$, will perform the feedback function. Therefore, the circuit in Figure 49-3 will oscillate if the following conditions prevail: the signal fed back to the grid is regenerative, the value of the capacitive reactance of the $C_{gp}$ is such as to aid in producing regenerative feedback, and the amount of regenerative feedback is sufficient to overcome resistive losses in the grid circuit. The circuit will oscillate at a frequency slightly below the resonant frequency of the tank circuits, and both tank circuits will appear as an inductive impedance to the oscillating frequency.

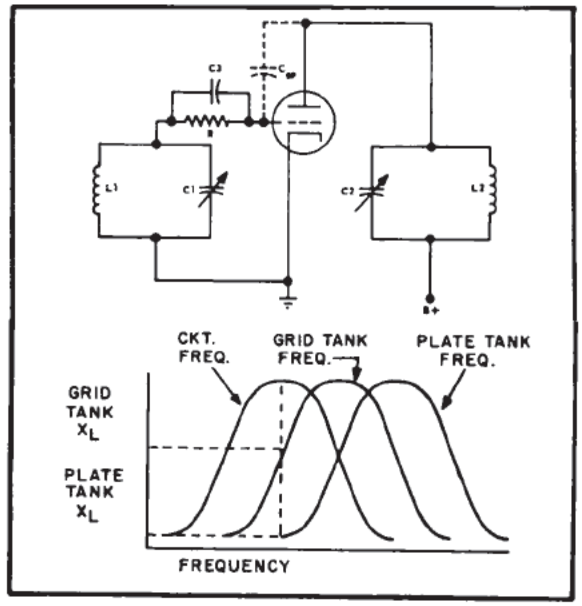


Figure 49-3 - TPTG oscillator circuit.

Before tuning the grid and plate tanks so that they present an inductive appearance, and discussing the importance of this in the operation of the circuit, an analysis of the circuit will be made with both tanks tuned to their natural resonant frequency. It will be assumed that the interelectrode capacitance($C_{gp}$) is very small, thereby, having a very large value of capacitive reactance. When tuned in this manner the circuit will appear as shown in the equivalent ac circuit, Figure 49-4A. Since the tanks are tuned to the operating frequency, they will appear resistive. The $X_c$ value for the grid-to-plate capacitance is higher than the impedance, $Z_g$, offered by the grid tank. In fact, it is usually designed to be ten times greater.

Figure 49-4B shows a vector diagram which indicates the phase relationship between the various ac voltages and currents in the circuit. It is assumed that the grid voltage vector ($e_g$) represents an initial signal applied to the grid. The plate current ($i_p$) is in phase with this grid voltage and the plate voltage ($e_p$) is 180° out of phase with this grid voltage as shown in the vector diagram.

A1. It multiplies either the local oscillator frequency of the receiver or the transmitter oscillator frequency. By using multiplication, a very stable oscillator may be used.

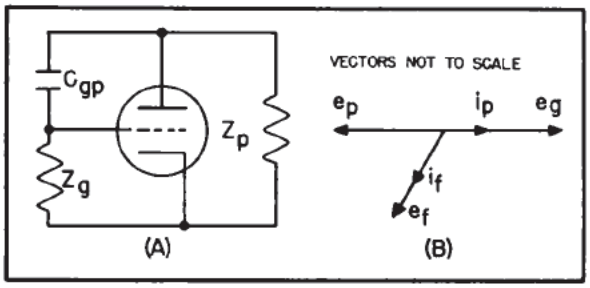


Figure 49-4 - TPTG oscillator equivalent circuit and vector diagram.

The plate voltage ($e_p$) is applied in parallel with the series combination of $C_{gp}$ and $Z_g$. The current flowing in this branch is termed the feedback current ($i_f$). Since the reactance of $C_{gp}$ is at least ten times greater than $Z_g$ this branch of the circuit will appear capacitive and the feedback current ($i_f$) will lead $e_p$ by an angle approaching 90°. The feedback current flowing through $Z_g$ will develope the feedback voltage ($e_f$). Since $Z_g$ is assumed to be resistive, $e_f$ will be in phase with $i_f$.

Figure 49-4B shows that the feedback voltage is not in phase with the grid signal, hence, oscillations will not occur. It is not necessary that $e_f$ be exactly in phase with $e_g$ in order for the circuit to sustain oscillations, however, the angular difference between the two should be small.

In order to oscillate, the tank circuits must appear slightly inductive, resulting in the ac equivalent circuit, Figure 49-5. The higher Q of the grid tank determines the frequency of the oscillator. Since the tank appears inductive it is represented in Figure 49-5 as a coil with

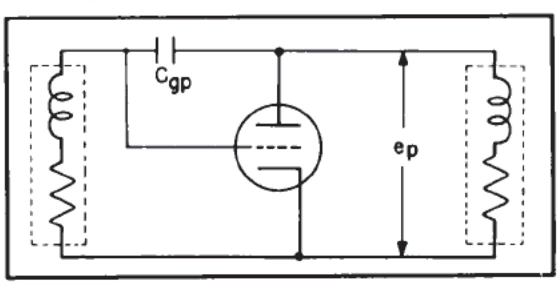


Figure 49-5 - TPTG oscillator showing both tank circuits tuned inductively.

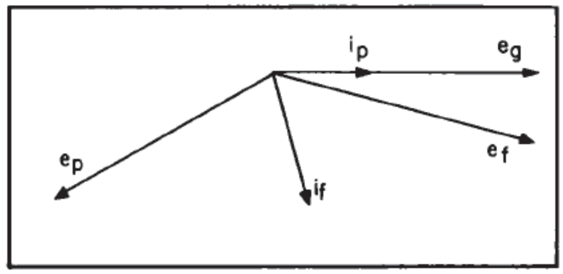


Figure 49-6 - TPTG Vector diagram, inductive load.

a series resistance. This resistance represents the effective resistance, or losses, of the tank. Normally the Q of the grid tank is greater than the Q of the plate tank. Therefore, the effective resistance in the grid circuit is smaller than the effective resistance in the plate circuit. Figure 49-6 is a vector diagram illustrating the relationship between the various currents and voltages in a TPTG oscillator when the tank circuits are tuned to appear inductive.

As before, $e_g$ is assumed to be a signal applied to the grid. Plate current ($i_p$) will be in phase with the grid voltage. Due to the 180° phase inversion introduced by the tube plus the inductive effect of the plate tank, will cause the plate voltage ($e_p$) to lead $e_g$ by some angle more than 180° but less than 270°. The more inductive the plate load impedance appears, the closer $e_p$ will be to leading $e_g$ by 270°.

The reactance of $C_{gp}$ is still much greater than the grid tank impedance, so the reactance of the series combination of $C_{gp}$ and the grid tank will remain predominately capacitive. Thus, causing the feedback current ($i_f$) to lead $e_p$ by an angle close to 90°. As shown previously, the feedback current will develope a voltage across the grid tank. However, it can be seen by the vectors in Figure 49-6, that since the grid tank is now inductive the feedback voltage ($e_f$) will lead $i_f$ by some angle approaching 90°. The net result of tuning the tanks to appear inductive is that the feedback voltage ($e_f$) is now nearly in phase with the grid signal and oscillations can be sustained. Proper adjustment of the tanks will bring $e_f$ and $e_g$ even closer to an in phase condition.

Tuning of the TPTG oscillator is accomplished by first adjusting the grid tank circuit to the APPROXIMATE desired operating frequency. This approximate grid tank adjustment is obtained through calibration marks on the tuning dial or operator experience. The plate tank is then adjusted until the circuit begins to oscillate. Once oscillations are obtained, the frequency of operation is compared with a frequency standard and the grid tank READJUSTED slightly to bring the circuit to the desired frequency of operation.

The plate tank is now READJUSTED to a frequency slightly below the frequency of operation of the circuit in order to insure more stable operation. Thus, the grid tank controls the frequency of operation while the plate tank determines the magnitude of the output signal.

Although it is possible to use the TPTG at relatively low frequencies its use is normally confined to the high frequency ranges by the value of the tubes interelectrode capacitance.

Q2. How must the plate tank be tuned to sustain oscillations?

Q3. If the triode were replaced by a pentode in the TPTG would the circuit still oscillate at the same frequency?

## 49-3. Crystal Oscillator

In order to obtain a higher degree of frequency stability in UHF transceiver oscillators, than is possible using a TPTG oscillator, a CRYSTAL OSCILLATOR is used. As stated previously, the operation of the crystal oscillator is very similiar to that of the TPTG. It will be shown, in this section, that a properly cut crystal possesses the characteristics of a resonant circuit and may, therefore, be used in place of a tuned circuit as a frequency controlling element. Before discussing the use and operation of a crystal in a circuit it will prove advantageous to discuss some of the properties and preparation of various types of crystals.

The control of frequency by means of crystals is based upon the PIEZOELECTRIC EFFECT. When certain crystals are compressed or stretched in specific directions, electric charges appear on the surface of the crystal. Conversly when such crystals are placed between two metallic surfaces across which a difference of potential exists, the crystals expand or contract. This interrelation between the electrical and mechanical properties of a crystal is termed the piezoelectric effect. If a slice of crystal is stretched along its length, so that its width contracts, opposite electrical charges appear across its faces, and a difference of potential is generated. If the slice of crystal is now compressed along its length, so that its width expands, the charges across its faces will reverse polarity. Thus, if alternately stretched and expanded a slice of crystal becomes a source of alternating voltage. Conversly if an alternating voltage is applied across the faces of a crystal slice, it will vibrate mechanically. The amplitude of these vibrations is very vigorous when the frequency of the ac voltage is equal to the natural mechanical frequency of vibration of the crystal. If all mechanical losses are overcome, the vibrations at this natural frequency will sustain themselves and generate electrical oscillations of a constant frequency. Accordingly a crystal can be substituted for the tuned tank circuit in an electron-tube oscillator such as the TPTG.

Practically all crystals exhibit the piezoelectric effect, but only a few are suitable as the equivalent of tuned circuits for frequency-control purposes. Among these are quartz, Rochelle salt, and tourmaline, of which Rochelle salt is the most active piezoelectric substance; that is, it generates the greatest amount of voltage for a given mechanical strain, However, the operation of a Rochelle salt crystal is affected to a large extent by heat, aging, mechanical shock, and moisture. Although these factors render this substance unsuitable for use in frequency control applications, it does find applications, as the active element in some microphones and phonograph pickups. Tourmaline is almost as good as quartz over a considerable frequency range, and is somewhat better than quartz in the range from 3 to 30 Mc, but it has the disadvantage of being a semiprecious stone. Its consequent high cost excludes it from general use. Quartz, although much less active than Rochelle salt, is used universally for frequency control of oscillators, because it is cheap, mechanically rugged, and expands very little with heat.

Crystals used in oscillator circuits are cut from natural or artificially grown quartz crystals which have the general form of a hexagonal prism, as shown in Figure 49-7A. They are rarely as symmetrical as that shown. Assuming, however, a symmetrical crystal, the cross section is hexagonal, as in part B and C of the Figure. The axis joining the points at each end, or apex, of the crystal is known as the optical or Z-axis. Stresses along this axis produce no piezoelectric effect. The X-axes pass through the hexagonal cross section at right angles to the Z-axis and are known as the ELECTRICAL AXES, because they are the directions of the greatest piezoelectric activity. The Y-axes, which are perpendicular to the faces of the crystal as well as to the Z-axis, are called the MECHANICAL AXES. A mechanical stress in the direction of any Y-axis produces an electrostatic stress, or charge, in the direction of that X-axis which is perpendicular to the Y-axis involved. The polarity of the charge depends on whether the mechanical strain is a compression or a tension. Conversely, an electrostatic stress, or voltage, applied in the direction of any electrical axis, produces a mechanical strain, either an expansion or a contraction, along that mechanical axis which is at right angles to the electrical axis. For example, if a crystal is compressed along the

A2. It must be tuned inductively to provide feedback of the proper phase.

A3. No. Because the value of the interelectrode capacitance in a pentode is much smaller. This smaller value of interelectrode capacitance will cause the circuit to oscillate at a higher frequency.

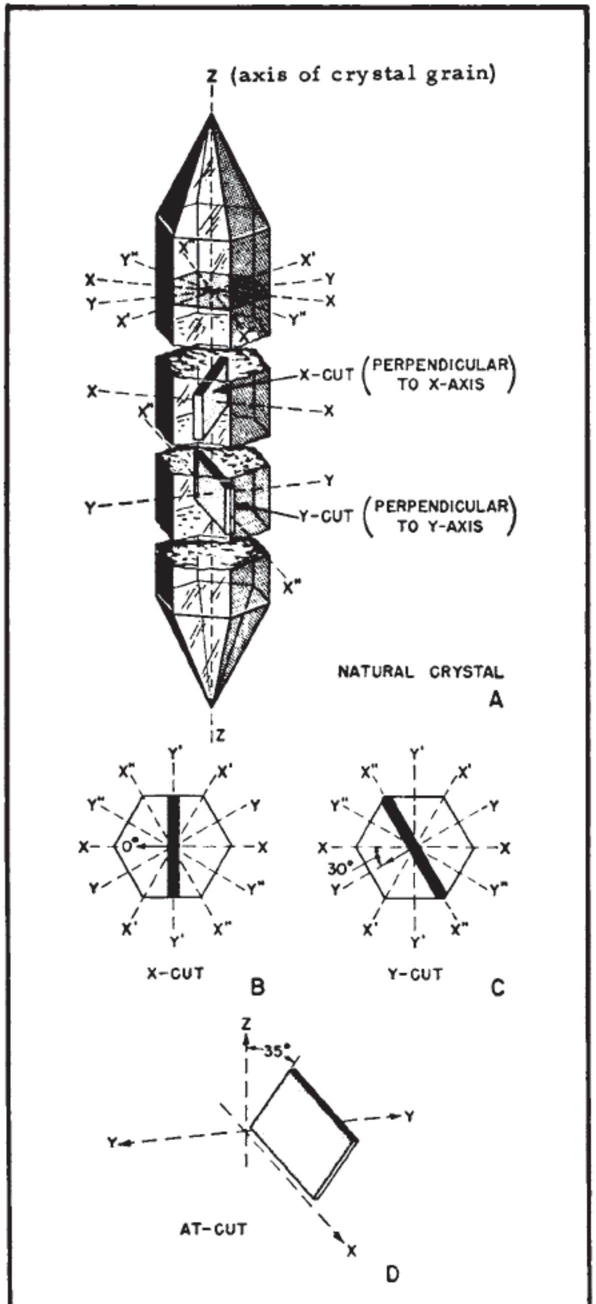


Figure 49-7 - Quartz crystal axes and cuts.

Y-axis, (Figure 49-7B), a voltage will appear on the faces of the crystal along the X-axis. If a voltage is applied along the X-axis (Figure 49-7C), of a crystal, it will expand or contract in the direction of the Y-axis. This interconnection between mechanical and electrical properties is exhibited by practically all sections cut from a piezoelectric crystal.

Crystal wafers can be cut from the mother crystal in a variety of directions along the axes. They are known as CUTS, and are identified by such designations as X, Y, AT, BT, GT, etc. Each has certain advantages, but, in general, one or more of the following properties are desired: ease of oscillation at intended frequency, a single frequency of oscillation, and minimum frequency changes resulting from temperature changes. The X-cut is sliced along a Y-axis and has its main parallel faces perpendicular to an X-axis, as in Part B of Figure 49-7. Although the diagram shows the section taken from the center of the crystal, the plate can be sliced from any part of the crystal, provided the orientation in respect to the axes is maintained. Part C illustrates the Y-cut, the long parallel faces of which are perpendicular to a Y-axis. This type is known also as a 30° cut, because the Y-axis passing through its center is at an angle of 30° in respect to the nearest X-axis. X and Y cuts have unfavorable temperature characteristics, as will be explained later. Better characteristics can be obtained by cutting plates at different angles of rotation about the X-axis. The Y-cut serves as zero-degree reference, since it is lined up with both X and Z-axes; that is, it lies in a plane formed by the X and Z-axes. Now, assume that the crystal wafer is rotated about the X-axis in a clockwise direction, so that it forms an angle of 35° with the Z-axis, as in Part D. The resulting slice is called the AT cut. Many other cuts exist, but they will not be discussed here.

Crystals used in oscillator circuits must be cut and ground to accurate dimensions. For instance, a typical quartz crystal resonant at 1 Mc would measure 1 inch by 1 inch by 0. 1125 inch. Crystals may be cut in various shapes, with crystals in the lower frequency range being square or rectangular and some of the crystals in the higher frequency ranges being disk shaped, similiar to a coin. For precise test work, crystals may be cut in the form of a flat ring. The type of cut also determines how active the crystal will be. Also, for any given crystal cut, the thinner the crystal, the higher the resonant frequency. The more care that is taken in cutting and grinding the crystal, the more accurate will be the final result. For example, crystals which are not ground to a uniform thickness may have two or more resonant frequencies. Usually one of these resonant

frequencies will be more pronounced than the others, with the less prominant ones being termed SPURIOUS frequencies.

The resonant frequency of quartz crystals is practically unaffected by changes in the load. Like most other materials, however, quartz expands slightly with an increase in temperature. This affects the resonant frequency of the crystal. The TEMPERATURE COEFFICIENT of the crystal refers to the increase or decrease in the resonant frequency, usually expressed in ppm (parts per million) or cycles per megacycle for an increase in temperature of $1^{\circ}C$. The temperature coefficient varies widely with different crystal cuts. The X- and Y-cuts have relatively high temperature coefficients at normal operating temperatures, with the coefficient of the X-cut being negative (frequency decreases as temperature increases) and the coefficient of the Y-cut being positive (frequency increases as temperature increases). The AT- and BT-cuts have relatively low temperature coefficients at normal operating temperatures, while the GT-cut has practically a zero coefficient over a very wide temperature range. The temperature coefficient also depends on the surrounding temperature at which it is measured.

Heating of the crystal can be caused by external conditions such as the high temperature of transmitter tubes and other components. Heating also can be caused by excessive RF currents flowing through the crystal. The slow shift of the resonant frequency resulting from crystal heating is called FREQUENCY DRIFT. This is avoided by use of crystals with nearly zero temperature coefficient, and also by maintaining the crystal at a constant temperature. To maintain the extremely close frequency tolerances required, the general practice is to construct the entire oscillator assembly in such a manner as to provide for nearly constant temperatures. This helps to avoid frequency drift resulting from contraction and expansion of circuit elements. The tube voltages are kept as constant as possible by suitable voltage-regulator circuits. In addition, the quartz crystal is operated in a constant-temperature oven. This oven is heated electrically and is held at constant temperature by special thermostats. The thermostats determine accurately any temperature variation and cause more or less current to flow through a heater element. The entire assembly usually is constructed of an aluminum shell inclosed by thick layers of insulating material to insulate the assembly. For extreme stability, the entire compartment can be placed inside still another temperature-controlled box. In this way, frequency stabilities as high as 1 part in 10,000,000 or better can be attained.

Crystals become practical circuit elements when they are associated with a crystal holder

Figure 49-8 - Crystal holder.

as shown in Figure 49-8. In a holder the crystal is placed between two metallic electrodes and forms a capacitor, the crystal itself being the dielectric. The crystal holder is arranged to add as little damping of the vibrations as possible, and yet it should hold the crystal rigidly in position. This is accomplished in various ways. In some holders, the crystal plate is clamped firmly between the metal electrodes; other holders permit an air gap between the crystal plate and one or both electrodes. The size of the air gap, the pressure on the crystal, and the size of the contact plates affect the operating frequency to some degree. The use of a holder with an adjustable air gap permits slight adjustments of frequency to be made. For the control of appreciable amounts of power, however, a holder which clamps the plate firmly is usually preferred.

At its resonant frequency, a crystal behaves like a tuned circuit so far as the electrical circuits associated with it are concerned. The crystal and its holder (shown in Figure 49-9A) can be replaced, therefore, by an equivalent circuit (as shown in part B of Figure 49-9). Here, $C_m$ represents the capacitance of the mounting, with the crystal in place between the electrodes but not vibrating. $C_g$ is the effective series capacitance introduced by the air gap when the contact plates do not touch the crystal. The series combination, L, R and C, represents the electrical equivalent of the vibrational characteristics of the quartz plate. The inductance, L, is the electrical equivalent of the crystal mass effective in the vibration. C is the electrical equivalent of the mechanical compliance (elasticity). R represents the electrical equivalent of the mechanical friction during vibration. The capacitance of the holder, $C_m$ is about 100 times as great as the vibrational capacitance, C, of the crystal itself.

The presence of both series and parallel resonant frequencies is revealed by the crystal resonance curve, in part C of Figure 49-9. This curve is very sharp, and extremely high Q's are easily attainable. It is found in practice that the $\frac{L}{C}$ ratio of the equivalent circuit is ex-

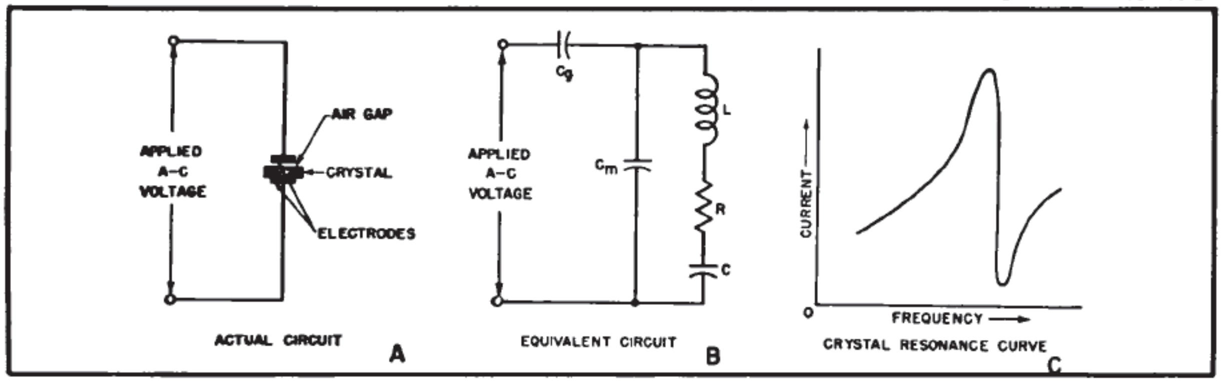


**Figure 49-9** - Equivalent circuit of crystal and mounting.

tremely large compared with that of a conventional tank circuit.

The circuit of a commonly used crystal oscillator, Figure 49-10, is seen to be the equivalent of the tuned-plate tuned grid oscillator, but with the crystal replacing the grid tank circuit. Feedback is obtained through the grid-to-plate capacitance, $C_{gp}$. The choke, RFC, keeps RF out of the grid-leak resistor, which provides the bias. The crystal and holder capacitances serve in the place of the grid leak capacitor. The crystal functions in the same manner as the grid tank circuit of the TPTG oscillator. It stores energy in mechanical form during one-half of the excitation voltage cycle, and releases it in electrical form during the second half of the cycle. The rate of storage and release of energy depends on the natural resonant frequency of the crystal and so determines the frequency of oscillation generated by the circuit. The losses in the crystal are overcome by the energy fed back through $C_{gp}$. The coupling between the tube and the crystal is determined primarily by the ratio $C/C_m$ (B of Figure 49-9), which was

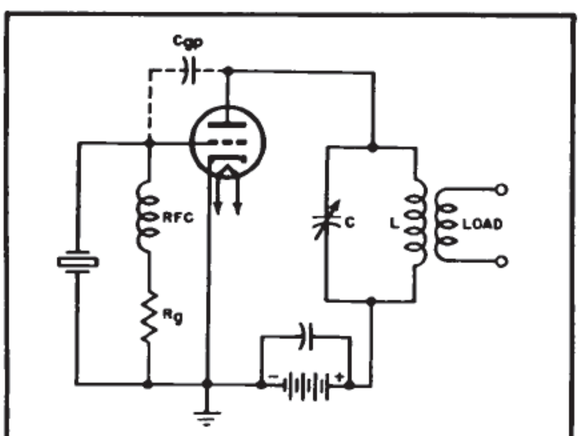


Figure 49-10 - Typical crystal oscillator.

seen to be very small. This small coupling, which is much less than in the TPTG oscillator, further improves frequency stability. As in the TPTG oscillator, the plate-tank circuit must be inductive, so that a negative resistance appears in the grid input circuit in parallel with the crystal to overcome the losses.

The resonance curve of the crystal (C of Figure 49-9) shown is obtained by tuning the plate circuit from a frequency below crystal resonance to one above crystal resonance. As the frequency is increased, series resonance of the crystal (mechanical resonance) is reached first, as indicated by the high crystal current peak in the resonance curve. The impedance is a minimum, and hence the current is a maximum for series resonance. In spite of the large current, oscillations cannot start because of improper phase relations. When the frequency of the plate tank is increased slightly above this value, parallel resonance of the crystal is reached: that is, the inductive reactance of the crystal proper equals the capacitive reactance of the crystal holder capacitance. This is indicated by the sharp drop of the crystal current in the resonance curve, and consequent high crystal impedance. Oscillations still cannot take place, however, since the plate tank is resistive at resonance. Another slight increase in plate tank frequency makes the plate circuit inductive, as required, and oscillations commence. Since this frequency is above crystal resonance, the equivalent crystal circuit is also slightly inductive. This inductive reactance is canceled out by the effective grid input capacitance, so that the entire grid circuit (crystal) plus tube input capacitance) is in parallel resonance. Maximum power output of the oscillator occurs at this frequency. If the frequency of the plate tank is increased still further, the plate tank impedance drops and less energy is fed back. Also, the increasing effective capacitance across the cry-

stal shunts the crystal and robs it of excitation. Under these conditions, oscillation stops. The presence of oscillations can be detected by a sharp drop in plate current as the frequency of the plate tank is raised.

Since the crystal will oscillate only at its resonant frequency, the frequency of oscillation remains constant at that value over a wide range of adjustment of tuning capacitor C. The power output changes substantially, however, when C is varied. Nevertheless, because of the shunting effect of the effective input capacitance in parallel with the crystal, the frequency of oscillations can depart slightly from the crystal resonance frequency for appreciable detuning of the plate tank without stopping the oscillations. For example, an increase of 2 percent in the tuning capacitance of the tank can reduce the frequency of oscillations by about 20 cycles per megacycle. Changes in plate voltage, filament voltage, and replacement of tubes also have very slight effects on the frequency of oscillation.

The outstanding characteristic of the crystal is the extreme sharpness of its resonance curve because of a very high effective Q. Because of this characteristic, the crystal will oscillate only over a very narrow frequency range, and, consequently, the frequency stability of a crystal oscillator is extremely high. These characteristics are utilized in military communications equipment where close frequency tolerances are required. Crystal oscillators are used not only in fixing the frequency of transmitters, but also as frequency standards for measurement purposes. If a low-frequency crystal is used in a circuit whose output is not tuned, a large number of harmonics are created. Therefore, a great number of calibration frequencies can be obtained with a single quartz crystal. A crystal-controlled oscillator is a fixed-frequency oscillator. A disadvantage is that a different crystal must be used for each desired frequency (or multiples of that frequency). In many applications, it is required to change the frequency of the transmitter rapidly and continuously. For these applications the ordinary variable-frequency oscillator is preferred, since it may be operated at any frequency within a band at the turn of a dial. Another limitation of the crystal oscillator is its relatively low power output. The power obtainable from a crystal oscillator is limited at low frequencies by the strains which the vibrations set up in the crystal structure. If the vibrations are too intense, they will crack the crystal. At high frequencies, the available power is limited by heating of the crystal. The heating is produced by the large RF crystal current necessary to obtain the excitation for substantial power output. Crystal heating short of the danger point results in frequency drift, thus lowering frequency stability. Excessive RF current can fracture the crystal. Operating a crystal up to its limits, 50 to 100 watts output may be obtained from a crystal oscillator circuit. It usually is preferred, however, to operate the crystal with a light load, and obtain the required output power through amplification in succeeding stages of the transmitter. In general, the crystal oscillator is considered as a frequency-generating device, with power output of secondary importance.

Q4. Why is a grid leak capacitor unnecessary in a crystal controlled oscillator?

## 49-4. Buffer Amplifier

As mentioned previously, a buffer amplifier is placed between the oscillator and the power amplifier to isolate the oscillator from the load thus improving the frequency stability of the oscillator. If the frequency of the plate tank circuit of the buffer amplifier is the same as that of the oscillator driving it, the stage is a conventional type of amplifier, usually class C.

Most of the transmitters operating in the UHF band are crystal controlled. However, crystals having a fundamental frequency in the UHF band are not practical because of physical restrictions, such as the difficulty in grinding crystals, and their extreme fragility. Therefore, the transmitter usually employs a low-frequency oscillator followed by a number of frequency multipliers. The number of multipliers used is dependent upon the frequency desired. Of course, the multiplier stages must be accurately tuned to the correct harmonic frequency. In this arrangement, the crystal is larger and more substantial than it would be if it were operated at the higher frequencies.

If the plate tank of the buffer amplifier is tuned to the second harmonic (in order to increase the frequency of the radiated signal) of the driving signal applied to the grid, the stage becomes a frequency doubler and the output voltage has a frequency equal to twice that of the input. Likewise, the buffer amplifier may become a tripler or a quadrupler.

A frequency-doubler stage is shown in Figure 49-11. The plate tank is tuned to twice the frequency of the grid tank. If $L_1$ is equal to 10 microhenrys and $C_1$ is equal to 25.3 picofarads, the resonant frequency of the grid tank may be found by the use of equation (11-16).

$$f_o = \frac{1}{2\pi\sqrt{LC}} \qquad (11\text{-}16)$$

$$f_o = \frac{1}{6.28 \times \sqrt{10 \times 25.3 \times 10^{-18}}}$$

$$f_o = 10 \text{ Mc}$$

A4. In the crystal controlled oscillator, the crystal holder acts as a capacitor in conjunction with the grid resistor to develop grid leak bias.

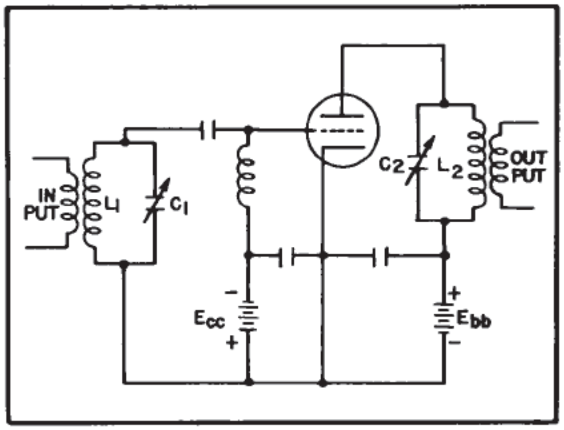


Figure 49-11 - Frequency doubler.

If the plate tank coil has an inductance of 10 microhenrys and the resonant frequency of the plate tank is 20 Mc, the plate tank capacitance, $C_2$, will be 6.25 picofarad.

The capacitance of $C_2$ may be verified by rearranging equation (11-16) and inserting values.

$$f_o = \frac{1}{2\pi \sqrt{LC}} \qquad (11\text{-}16)$$

expanding:

$$f_o = \frac{1}{2\pi} \times \frac{1}{\sqrt{LC}}$$

simplifying by the reciprocal of $2\pi$:

$$f_o = \frac{0.159}{\sqrt{LC}}$$

solving for C:

$$C = \frac{(0.159)^2}{f^2 L}$$

from which

$$C = \frac{(0.159)^2}{(20 \times 10^6)^2 \times 10 \times 10^{-6}}$$

$$C = 6.25 \text{ pf}$$

The plate voltage, grid voltage, and plate current waveforms are shown in Figure 49-12. The dotted lines indicate operation as a class C amplifier without frequency multiplication, and the solid curves indicate operation as a frequency doubler.

When the plate tank is tuned to the frequency of the driving signal, it appears as a high impedance to the signal and a large portion of the B+ voltage is dropped across the tank; this results in a low plate current. As the plate dissipation in heat is a function of plate voltage and plate current, the heat dissipated by the plate is a relatively small amount.

However, when the plate tank is tuned to a harmonic of the driving signal, it appears as a low impedance to the signal and a relatively small portion of the B+ voltage is dropped across

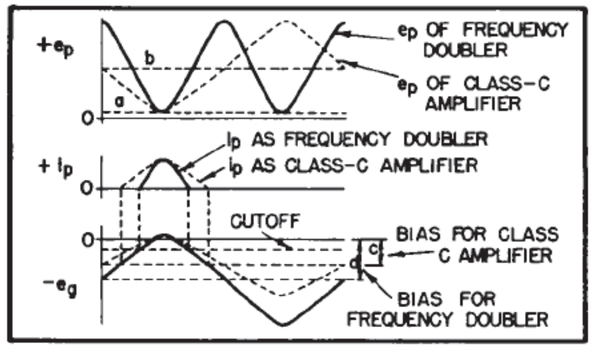


Figure 49-12 - Frequency doubler waveforms.

the tank circuit; this results in a high plate voltage and a high plate current during the time the grid is above cut-off. Thus, a large amount of power is dissipated by the plate as heat.

The higher the harmonic, the lower the impedance and the more power loss as heat.

The plate dissipation loss is kept within the rating of the triode by increasing the operating bias, thus reducing the length of time that plate current flows.

In every case it is necessary to increase the operating bias and the grid driving signal as the frequency multiplication increases in order not to overheat the triode plate. The flywheel effect in the plate tank supplies the missing cycles of grid drive and the output is approximately an undamped wave having sine waveform.

Three important conditions must prevail in order to obtain frequency multiplication; high grid bias, tunable plate tank, and high grid-driving voltage. If the second harmonic is selected, the stage is called a frequency doubler; if the third is selected, it is called a frequency tripler, and so forth.

Certain amplifier circuits are suited to the generation of even harmonics and others to the generation of odd harmonics. Push-pull amplifiers produce only odd harmonics multiplication— third, fifth, seventh, and so forth. If the grids of two triodes are connected in push-pull and the plates in parallel as shown in Figure 49-13, even order harmonics may be produced.

The grid signals are 180° out of phase. When one grid voltage is maximum positive, the other

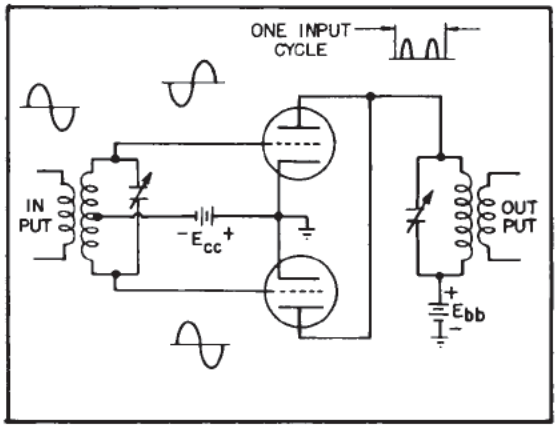


Figure 49-13 - Even-order harmonic frequency multiplier.

is maximum negative, and the second alternation of the cycle reverses the respective potentials. Thus pulsating plate current flows first in one tube and then in the other. Because the plates are connected in parallel, the output pulses are in the same direction and the plate tank circuit receives two pulses for each input on the grids. This type of doubler is capable of greater output and higher plate efficiency than the single-tube type.

Q5. Why is it necessary to increase bias when increasing frequency multiplication?

## 49-5. Power Amplifier

UHF is normally low power transmission. For that reason it is not necessary to use a high-power capacity tube in the output. In some cases, the final-power output tube is also used as a multiplier tube. Figure 49-14 shows a UHF power amplifier which serves as both push-pull output tube and frequency tripler. Two beam power tubes are used in the push-pull, plate-and-screen modulated final power amplifier and frequency tripler circuit. The grid tank circuit of this amplifier is composed of $L_{116}$, $L_{117}$, $C_{178A}$, and $C_{179}$. The tank circuit is provided additional values of capacitance and inductance by stray inductance, stray capacitance, and the input capacitance of the tubes. Capacitor $C_{178A}$ is a split-stator capacitor and is ganged tuned with the four sections of $C_{145}$ and with $C_{178B}$. Capacitor $C_{179}$ is also a split stator capacitor and is used for alignment adjustments. The input signal is applied through a coaxial cable and the low impedance link, $L_{115}$.

The plate tank circuit is composed of a parallel transmission line inductance and two ganged parallel split capacitors. This plate tank assembly is designated as $Z_{102}$ and is actually a TUNED LINE (to be discussed shortly). A movable shorting bar across one half of the plate tank inductor is used to obtain the proper tuning range. The plate circuit is tuned to the third harmonic of the input signal, covering the range of 225 to 400 Mc. The modulated output of this stage is fed to the antenna relay and antenna system through a filter network.

Plate voltage is applied to the power amplifier tubes ($V_{117}$ and $V_{118}$) through the secondary winding of the modulation transformer, and the RF filter composed of $L_{118}$, $C_{172}$, $C_{177}$, and $C_{182}$. Screen-grid voltage is applied to the screen through the secondary winding of the modulation transformer. Capacitor $C_{175}$ is a feed-through bypass capacitor which helps keep the screen lead at RF ground potential.

The output filter prevents undesired frequencies developed in the frequency multiplier from reaching the antenna and being radiated. The filter provides a satisfactory load impedance in the operating band of 225 to 400 Mc, but at spurious frequencies its impedance is low and the undesired signals are bypassed to ground. The output filter is tuned through the output frequency band by two split-stator capacitors, $C_{178B}$ and $C_{178C}$. Capacitor $C_{178B}$ is gang

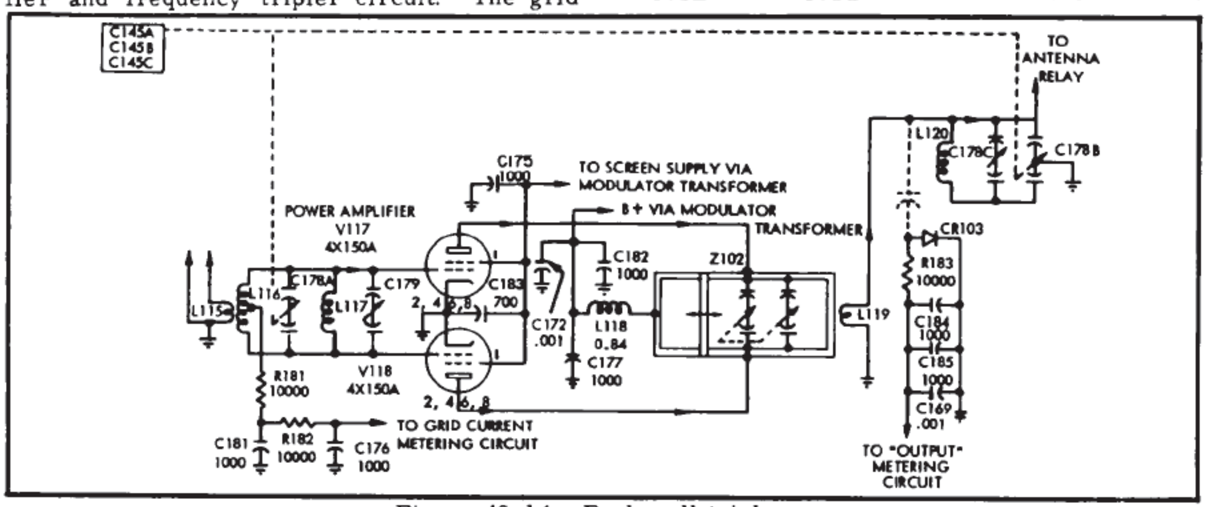


Figure 49-14 - Push-pull tripler.

A5. To prevent overheating the triode plate.

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tuned with the previous stages and $C_{178C}$ is used as a trimmer. Power input to the filter and antenna is through pickup link $L_{119}$, a rectangular loop inductively coupled to tank assembly $Z_{102}$ of the power amplifier. Inductance $L_{120}$ is the tank inductance in the output filter.

The output metering circuit is used to determine the output power.

In the circuit of Figure 49-14, a tuned line was used to replace the tank circuit in the frequency multiplier. The use of a tuned line is conventional in UHF systems. In fact, there are a number of oscillators which use the tuned line. The reason for this is that the inductive and capacitive elements would become too small (due to the high operating frequencies) to be practical. In addition, the skin effect in the coils introduces ac resistance which reduces the Q of the oscillator. To avoid these defects, the tuned circuit for UHF oscillators usually consist of a quarter wavelength transmission line called a tuned line. A segment of transmission line one-quarter wavelength long and shorted at the receiving end has the characteristics of a parallel resonant circuit. This action can be analyzed by referring to the quarter-wave shorted transmission line segment shown in Figure 49-15A. At the shorted end, the current will be maximum and the voltage will be minimum. The maximum current condition is characteristic of a series resonant circuit.

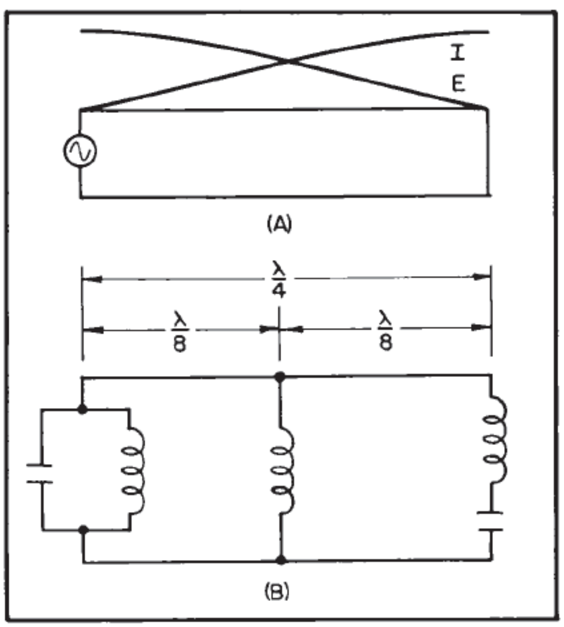


Figure 49-15 - Shorted quarter-wave tuned line.

Part B of Figure 49-15, shows the equivalent circuit conditions of the quarter-wavelength line at various points between the shorted end and the source. At the shorted end part B shows the line to be equal to a series resonant circuit. One-eighth wavelength back toward the source, at the point where the current and voltage waveforms cross, the circuit will appear inductive. This is shown in the equivalent circuit by the inductance connected across the line. The circuit will appear inductive in varying degrees between the source and the shorted end. At the source, the current will be minimum and the voltage will be maximum—a condition characteristic of a parallel resonant circuit. Since the parallel resonant circuit is the first element that the source sees, it may be said that the shorted quarter wave line will act like a parallel resonant circuit as far as the source is concerned. The tuned lines (mentioned previously as a replacement for a UHF tank circuit) are adjustable to facilitate a change in their operating frequency. This adjustment is made by means of a shorting bar, shown in Figure 49-16. The function of the shorting bar is to change the

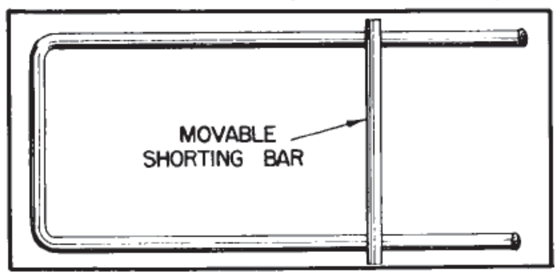


Figure 49-16 - Tuned line with shorting bar.

physical length of the line. This will, of course, change the wavelength of the line and thereby the frequency at which the line displays parallel resonant characteristics to the source. There-

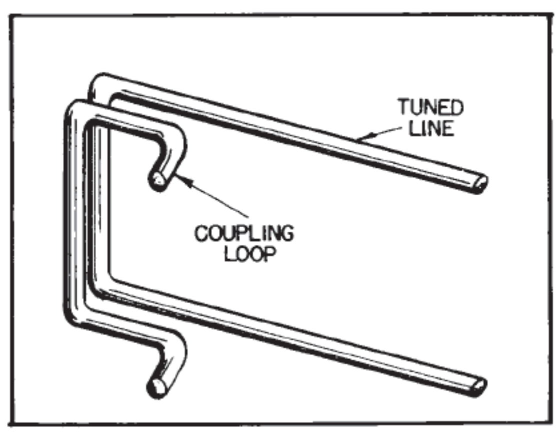


Figure 49-17 - Loop Coupling.

fore, assuming the tube to be the source, connecting tuned lines between plate and cathode and grid and cathode and adjusting the shorting bars for the proper frequencies will cause the tube to "see" resonant tanks in its plate and grid circuits.

Coupling from the plate circuit tuned line may be accomplished by various methods. One of the most popular methods is called LOOP coupling. Loop coupling is shown in Figure 49-17. The loop is merely a conductor placed in position with the tuned line so that mutual inductance exists between the two.

Q6. Why is it practical to replace the conventional tank circuit with a tuned line at UHF frequencies?

## RF AMPLIFIERS

At UHF frequencies conventional electron tubes possess disadvantages that decrease their efficiency of operation considerably, in many cases rendering them unusable in certain applications. These disadvantages fall into certain categories, such as: limitations imposed by the physical structure of the tube, limitations imposed by radio-frequency losses, limitations imposed by TRANSIT TIME, and the limitation imposed by the production of noise. It is the purpose of the following sections to discuss the cause, and in some cases the curve, of these limitations in UHF amplifiers.

### 49-6. Limitations Imposed by Tube Structure

At ultra-high frequencies the electron tube must be viewed as an ac circuit element consisting of interelectrode capacitances and inductances which are inherent in the structure of the tube. When the electron tube is used in an application where the grid or plate circuits contain resonant circuits, the resultant action of the interelectrode capacitances is to increase the tank capacitance. At lower operating frequencies where the tank capacitance is relatively large, this effect is negligible. However, as the operating frequency is raised into the UHF range, the interelectrode capacitances become a proportionately greater part of the tank capacitance. Therefore, changes in tube capacitance (due to heating, Miller effect, loading, aging, etc.) will effect the tuned circuits resonant frequency.

Another limitation of the conventional tube at UHF frequencies is the inductance of the elements themselves and their lead wires. At low operating frequencies it is common practice to assume that a straight piece of wire exhibits practically zero dc and ac resistance.

For example: a straight piece of wire 0.04 inch in diameter and 4 inches long, possesses an inductance of approximately 0.1 microhenry. When used at an ordinary broadcast frequency of 1 Mc, this wire will exhibit a inductive reactance of:

$$X_L = 2\pi f L$$

$$X_L = 6.28 \times 1 \times 10^6 \times 0.1 \times 10^{-6}$$

$$X_L = 0.628 \text{ ohms}$$

This amount of reactance is negligible. However, this same piece of wire used in a circuit at a UHF frequency of 300 Mc possesses an inductive reactance of:

$$X_L = 2\pi f L$$

$$X_L = 6.28 \times 300 \times 10^6 \times 0.1 \times 10^{-6}$$

$$X_L = 188.4 \text{ ohms}$$

This amount of reactance will exhibit considerable choking effect. In addition, the inductance of the cathode lead is common to the grid and plate circuits, so that it provides degenerative feedback.

Low interelectrode capacitance can be attained either by reducing the electrode size or by spreading the electrodes farther apart. However, unless abnormally high voltages can be used, spreading the electrodes will have the undesired effect of increasing the electron transit time. If all the linear dimensions of an electron tube are reduced, lead inductance and transit time will be REDUCED. However, reduction of the physical size of the tube reduces its power-handling ability, since only small areas are present for dissipating heat.

In addition to the reduction that results from decreasing the physical size of the tube, the lead inductances may be further decreased by making the leads of large diameter rods and of the shortest length that will provide a safe insulating distance between the anode and other tube terminals. Thus the leads in most ultra-high frequency tubes are brought out straight through the tube envelope, and no conventional tube base is used.

Q7. Would the interelectrode capacitances of a conventional electron tube exhibit lower or higher reactances if used in a UHF circuit?

### 49-7. Limitations Imposed by RF Losses

As the frequencies of an oscillator is raised the RF circuit losses increase. These losses are due to many factors, among which is the

A6. The size of the tank circuit components at UHF frequencies are physically impractical.

A7. Lower. $X_c$ decreases as frequency increases.

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increase of skin effect. Skin effect forces current to flow in a thin layer on the surface of the conductor. The higher the frequency, the thinner will be the layer in which the current flows. The $I^2R$ losses that take place in a given conductor must increase, then, as the frequency increases, since the area in which current can flow is less, making the resistance greater. Skin effect is reduced by using conductors of large diameter so that the current can flow through a reasonably large cross-sectional area even though the depth of penetration is small. Thus, using large leads for ultra-high-frequency tubes reduces not only the lead inductance but also the lead resistance. As a means of further reducing skin effect, conductors are often plated with a low-resistivity metal such as silver. In cases where corrosion could convert the surface of the conductor to a high-resistivity oxide, the conductor is plated with gold which does not corrode and which has excellent conductivity.

Because the reactance of interelectrode capacitance and distributed capacitance becomes small for UHF frequencies, the magnitude of the charging current for these capacitances become large. Such currents contribute nothing to the power output, but in flowing through the resistance of the circuit they do produce losses. Because of skin effect, these currents follow the surface of the metal electrodes. In some cases this may cause excessive localized heating at the junctions of the electrodes and the glass envelope, which may result in cracked seals and failure of the tube.

In oscillators for ultra-high-frequency use, not only must the tube losses be kept very low, but the losses in the associated circuits must also be kept as low as possible. For this reason, the tuned circuits associated with ultra-high-frequency oscillators are usually resonant sections of transmission line rather than coil and capacitor combinations.

The Q of a quarter-wave short-circuited section of line can be made much higher than that of a conventional tank circuit because it is feasible to make a tuned line of conductors of larger diameter than is possible with the conventional inductor, making the skin effect less. In addition to their low losses, tuned transmission lines are used as circuit elements in ultra-high-frequency oscillators because the tube leads may act as extensions of the transmission line. Thus the interelectrode capacitances and lead inductances are incorporated as part of the tuned circuit.

## 49-8. Limitations Imposed by Transit Time

At low radio frequencies transit time is negligible, since it occupies only a comparitively small portion of an oscillatory period. But as the frequency of operation becomes higher, transit time occupies an appreciable portion of this period and produces undesirable effects in tube operation.

In low-frequency operation it is usually taken for granted that electrons leaving the cathode reach the plate instantaneously. Although nothing in nature happens instantaneously, no harm is done by this assumption, so long as the actual time of flight of electrons between the cathode and plate is negligible compared to the duration of one cycle. For example, transit time of one-thousandth of a microsecond ($10^{-9}$ second), which is not unusual, is only one-thousandth of a cycle at a frequency of 1 Mc. However, the same transit time becomes one-tenth or a greater part of a cycle if the frequency is 100 Mc or higher.

It has been found experimentally that with the total transit time less than one-tenth of a cycle the tube operates satisfactorily. At transit times longer than one-tenth cycle the efficiency drops considerably. When the transit time approaches a quarter of a cycle at the oscillating frequency the tube usually does not oscillate at all. This decrease in output is caused, in part, by the shift in phase between the plate current and the grid voltage and the decrease of the effective resistance between grid and cathode, which results from the relatively long transit time.

The effect of transit time is of special concern in connection with the input impedance of the tube. Part of the current that flows in the grid circuit is the current which charges the grid-to-plate capacitance, $C_{gp}$. The voltage that produces this current is the vector sum of the input voltage (grid-to-cathode) and the output voltage (plate-to-cathode). At lower frequencies with a resistive load, these two voltages are $180^\circ$ out of phase and add algebraically to determine the charging current. This current is $90^\circ$ out of phase with the input voltage. However, at higher frequencies, where transit time is an important factor, the plate current begins to lag the input voltage. This is caused by the fact that it takes an appreciable part of a cycle for the electrons to travel from cathode to plate, the current which is arriving at the plate at any instant is different from the current leaving the cathode at that same instant. Assuming that a given number of electrons start toward the plate at the same instant, they will not arrive at the

plate until a short time later. During this period of time the phase of the input voltage will have changed. Since the electrons which actually strike the plate make up the plate current, it can be seen that the plate current will lag the input voltage and the current emitted from the cathode. As a result of this lag the power output decreases and the plate dissipation increases. A further result of this lag will be a consumption of power in the input circuit due to an induced grid current. This induced grid current is caused in the following manner.

When an electron, which is a negative charge, approaches an electrode, it induces a positive charge on the electrode. As the electron approaches, the positive charge flows to the electrode; as the electron recedes, the positive charge flows away from the electrode. Thus, the electrons that form the plate current in a vacuum cause electrostatically induced currents in the grid as they move past it. In a low-frequency oscillator in which the grid is negative and the transit time is negligible, the number of electrons approaching the grid is always the same as the number of electrons going away from the grid. The current induced on one side of the grid by the approaching electrons is equal to that induced on the other side by the receding electrons. Since these currents are in opposite directions, the combined effect is zero.

However, when the transit time of the oscillator is an appreciable part of a cycle, the number of electrons approaching the grid is not equal at all times to the number going away. As a result, the electrostatically induced currents do not cancel. Thus, grid current can flow in an ultra-high-frequency oscillator EVEN WHEN THE GRID IS NEGATIVE RELATIVE TO THE CATHODE. This current consists of a movement of positive charges back and forth in the grid structure. The effect of this current is to produce in the grid itself a loss which may be considered as a loss that takes place in an imaginary input resistor connected between grid and cathode. The resistance of this imaginary resistor decreases rapidly as the frequency rises. At ultra high frequencies this resistance may become so low that the grid is practically short-circuited to the cathode, preventing proper excitation of the tube. This additional ultra-high-frequency grid loss raises its temperature, which may become another limitation on the frequency at which a tube may generate oscillations.

Transit time may be decreased by reducing the spacing between electrodes, or by increasing the electrode voltages. Since insulation considerations prevent raising voltages greatly in many applications, tubes of special design are usually employed in applications where transit time effects may become serious.

Q8. What part of a period would a transit time of one-thousandth of a microsecond ($1 \times 10^{-9}$) be at an operating frequency of 400 Mc?

## 49-9. Limitations Imposed by Noise

There are several sources of noise in the UHF RF amplifier circuit. Noise is not only generated by the amplifier tube, but also within the conductors which form the circuit. This latter noise source is due to thermal agitation of electrons within the conductor. This type of noise is also referred to as RESISTANCE NOISE, CIRCUIT NOISE, or THERMAL NOISE. This noise is generated by fluctuations in circuit current caused by imperfections in the crystal lattice structure of the conductors.

Tube noise is a result of irregularities in electron flow through the tube. Since there are several sources of noise, they will be discussed individually.

SHOT EFFECT: is a source of noise within a tube and is due to an irregularity in plate current, caused by variations in emission of electrons from the cathode. The cathode of an electron tube is coated with an emitting material. It is possible that the coating may not be spread uniformally over the surface of the cathode. It is also possible for impurities to be mixed in with the emitting substance. Both of these conditions will produce variations in the rate of emission of electrons from the material, thus generating noise. The noise generated by shot effect is greatly reduced by the presence of a space charge, which acts in such a manner as to absorb any instantaneous variations in emission.

PARTITION NOISE: occurs because of chance variations in the division of current between two or more positive electrodes. For example, partition noise is found in triodes when the control grid is positive and current divides between grid and plate. Although the grid current will only be a small portion of the total tube current, at any one instant there will be more or less grid current flow, thereby, causing a variation in the division of current between electrodes. Use of multigrid tubes causes an increase in partition noise because there are more elements for the current to divide between.

INDUCED GRID NOISE: is produced as a result of variations in the electron stream passing adjacent to the grid (as explained in section 49-8). The noise is produced by the movement of charges in the grid structure as electrons approach and then recede from the grid. This type of noise will occur in every tube except the diode.

GAS NOISE: is generated by the random rate of production of ions due to collision. This type of noise is minimized by the use of high vacuum tubes.

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A8. 0.4 of a period. First determine the period of 400 Mc signal ($T = \frac{1}{t} = \frac{1}{4 \times 10^8} = 2500$ picosecond.) Then determine what part of a period by dividing the transit time by the period.

$$\frac{1 \times 10^{-9}}{25 \times 10^{10}} = 0.4 \text{ of a period.}$$

SECONDARY EMISSION NOISE: is a result of the bombardment of the plate by high velocity electrons. Electrons knocked loose from the plate by this bombardment are called secondary emission electrons. When these electrons return to the plate they cause secondary emission noise.

FLICKER EFFECT: is caused by a low-frequency variation in emission that occurs with oxide-coated emitters.

Shot effect noise, partition noise, and induced grid noise are the most common and the most troublesome. It should be noted that all of these noises are caused by variations in electron flow.

To minimize the effects of transit time and noise generation, special tubes are used. It was mentioned that one of the effects of transit time is a high value of grid loss. These special tubes will also minimize this loss. The requirements of these types of tubes are as follows: they must have a short transit time, the inductance of the connecting leads and the capacitance between the tube electrodes must be very small. The area of the electrodes must also be small to decrease the interelectrode capacitance. The spacing of the electrodes is close, this decreases the transit time of the tube. A relatively high voltage applied to the electrodes also helps minimize the transit time. The inductance of the connecting leads is reduced by the use of short leads of large diameter, or by the use of multiple leads.

Because of the electrode spacing, if a great deal of power output is expected; a high plate loss should be anticipated. There are several tubes that have the characteristics just mentioned. Two of these are the acorn tube and the lighthouse tube, shown in Figure 49-18.

The acorn tube received its unusual name because of its acorn like shape. It is characterized by the use of short leads which come directly through the tube envelope instead of the base of the tube. The purpose of this arrangement is to minimize lead length. Figure 49-18 shows a triode and a pentode acorn tube. The lighthouse tube features a system designed to place the electrodes as close to each other as possible. This tube is particularly suited for use with cavity resonators or with coaxial resonators. This feature is embodied in the structure because of the closely spaced cylindrical construction of the device. This cylindrical structure permits the handling of greater amounts of power.

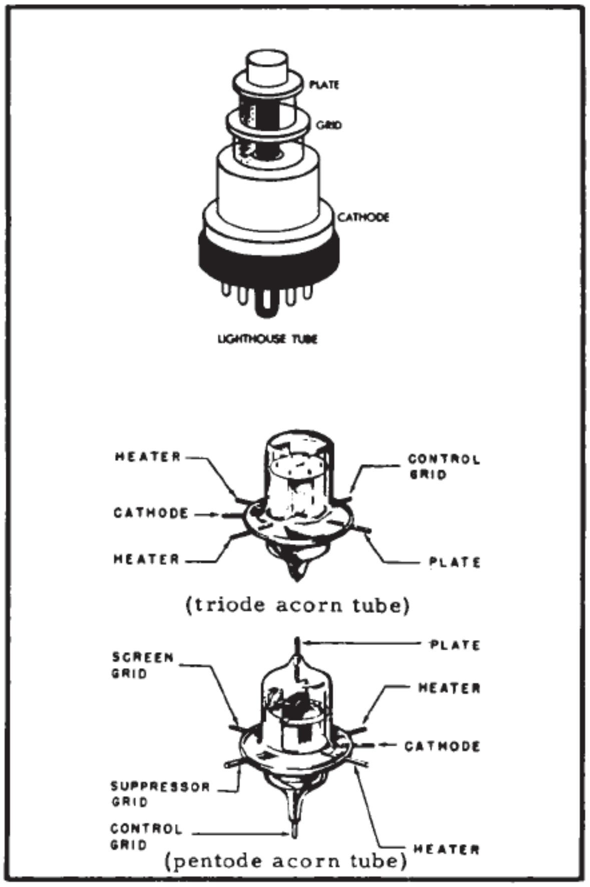

Figure 49-18 - Low-transit-time tubes.

The signal-to-noise ratio of a receiver is primarily determined in the first stage. Therefore, when an RF amplifier is used it must possess a very high signal-to-noise ratio. To imporve the signal-to-noise ratio a special type of circuit is sometimes employed. This circuit is called a GROUNDED GRID AMPLIFIER. Normally, with the gain requirement expected from a first stage, it might be expected that a pentode would be used. However, due to the noise limitations discussed previously a pentode is not normally used. The grounded grid amplifier employs a triode tube.

The basic grounded grid amplifier is shown in Figure 49-19A. It is a grounded grid because, due to efficient bypassing of the grid bias supply, the grid is at ac ground potential.

The input for the circuit is shown in the cathode circuit. The equivalent circuit for the

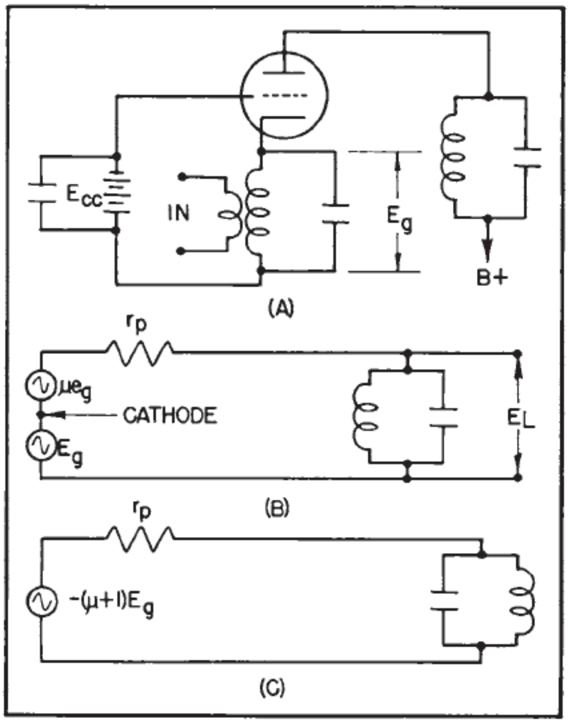


Figure 49-19 - Grounded grid amplifier.

grounded grid amplifier is shown in Figure 49-19B. It shows the input signal in series with the output of the tube for this reason, the amplified output $E_L$ is the same as though the grid were excited by the input signal. The amplification factor for that reason is $(\mu + 1)$ instead of simply $\mu$. Because the signal source is in series with the plate circuit it also supplies a portion of the energy dissipated in the plate circuit. One other characteristic of this type of amplifier is the relatively low input resistance. Using one generator the equivalent circuit appears as shown in Figure 49-19C. One advantage of this type of circuit arrangement is that neutralization is not necessary. The reason for this is that the grounded grid acts like a shield. As a result, the energy transfer between cathode and ground and plate to ground circuits (through the tube capacitances) is avoided. Due to the grounding of the grid the internally generated noise of this circuit will be reduced, since there will be a minimum grid induction effect.

Q9. What is induced grid noise?

## OSCILLATOR

### 49-10. UHF Transceiver Oscillator

Operation of an oscillator at UHF frequencies has been discussed at great length in previous sections of this chapter. Therefore, only a brief discussion of the oscillator in relation to the UHF transceiver will be presented here.

The oscillator used in the UHF transceiving must be highly stable in frequency. For this reason a crystal oscillator or a crystal CONTROLLED oscillator is used. The word controlled is emphasized because in some cases the crystal is not used as the resonant tank circuit, but rather as an additional device to accurately control the oscillations of a conventional LC tank. This control is accomplished, in most cases, by inserting the crystal in the feedback path and utilizing its series resonant characteristics. Thus, only energy at the exact series resonant frequency of the crystal will supply regenerative feedback to the oscillator circuit and strict control of the operating frequency can be maintained. Most transceivers have several crystals which may be inserted into the circuit by switching action, thereby changing the frequency of the oscillator circuit.

## MIXER

### 49-11. UHF Transceiver Mixer

The function of the mixer in the receiver section of a transceiver is the same as for any other receiver. In other words, the mixer provides a heterodyning action between a modulated RF signal and a constant amplitude oscillator signal, in order to produce a modulated IF signal.

It is desirable to have a relatively low value of IF frequency. In order to convert the incoming UHF signal (300 Mc for instance) to an IF frequency of 455 kc in one step, would require that the local oscillator have virtually perfect frequency stability. Otherwise, the slightest drift in the oscillator would shift the IF frequency a considerable amount. For this reason UHF transceivers usually convert from the incoming signal to the final IF frequency in two or more steps. This is termed DOUBLE CONVERSION, TRIPLE CONVERSION, etc. Double conversion is shown by use of the block diagram in Figure 49-20. The output of the RF amplifier, along with the signal from the

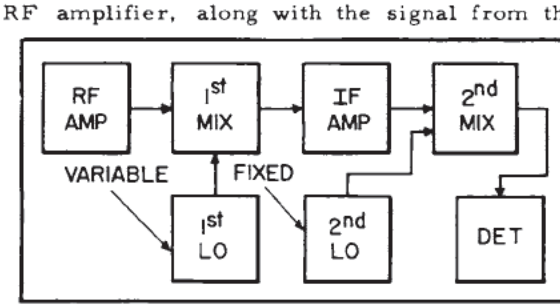


Figure 49-20 - Block diagram of double conversion.

A9. It is the result of electron movement in the grid structure, caused by cathode emitted electrons approaching, passing, and receding from the grid.

first local oscillator, is fed to the first mixer. The first local oscillator is made variable in order to provide tracking between the oscillator and the input signal. The output of the first mixer is a modulated IF frequency, which is then used as the RF input to the second mixer. Heterodyning of this signal with that of the second local oscillator will produce a lower IF frequency signal, which in this case is the final IF frequency. If a lower final IF frequency were desired, a third conversion system would be added merely by adding a third mixer and local oscillator. Notice that the second oscillator operates at a fixed frequency, as would any additional local oscillators. It is not necessary for any but the first oscillator to be variable since the output of the first mixer is a constant frequency.

## 49-12. IF Amplifier

It was previously mentioned that the IF amplifiers were responsible for the receiver gain and bandwidth. The IF amplifier is a high-gain circuit commonly employing pentode tubes. This amplifier is tuned to the frequency difference between the local oscillator and input frequencies. IF amplifiers may be composed of one or more stages of amplification depending on the amount of gain required. Figure 49-21 shows an IF amplifier stage.

Cathode bias is established by means of $R_1$ and $C_1$. Automatic volume control (avc) is applied to the grid of the tube through the secondary of the input IF transformer. In the IF stage, of a UHF receiver, the inductances (transformers, chokes, etc.) are resonated with the distributed and interelectrode capacitances.

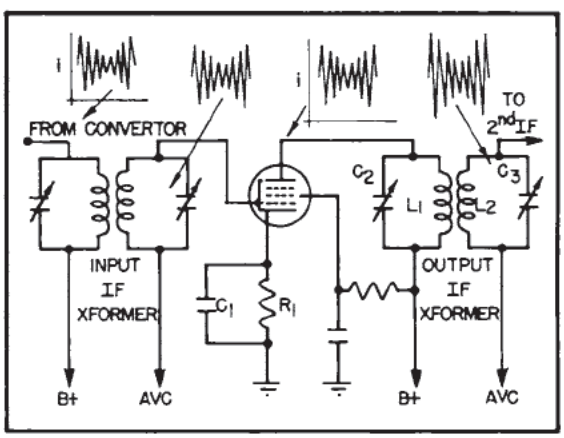


Figure 49-21 - IF amplifier stage.

## 49-13. Detector

Most superheterodyne receivers employ a diode as the detector. This type of detector is practical because of the high gain as well as the high selectivity of the IF stages. The diode detector has good linearity and can handle large signals without overloading. For reasons of space and economy, the diode detector and first audio amplifier are often included in the same envelope in modern superheterodyne receivers.

A simple diode detector is shown in Figure 49-22. The rectified voltage appears across $R_1$, which also serves as the volume-control potentiometer. Capacitor $C_2$ bypasses the RF component to ground, and $C_3$ couples the output of the detector to the first audio amplifier stage. The tuned circuit $L_2C_1$ is the secondary of the last IF transformer.

The time constant of $R_1C_2$ is long compared to the time for one IF cycle but short compared to the time for one AF cycle. In order to increase the high frequency response of the diode detector the time constant of $R_1C_2$ is reduced.

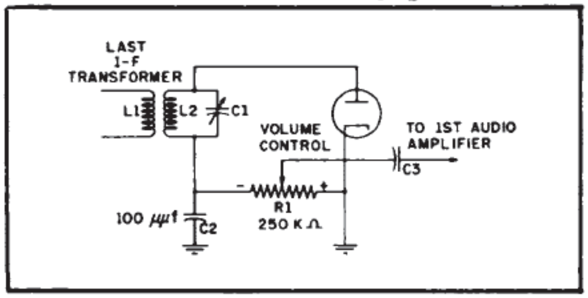


Figure 49-22 - Diode detector.

Diode load filter capacitor $C_2$ cannot discharge rapidly enough to follow modulation frequencies higher than 10,000 cps (in this case), and clipping results with all higher audio frequencies.

## 49-14. Audio Amplifier

The output from the detector stage is an audio frequency of a very low level. To increase the level of the audio, the output of the detector is then sent to an audio amplifier. The audio amplifier used in this receiver is similar to any audio amplifier. It should amplify equally all those frequencies in the audio range. Military receivers, however, are not interested in high fidelity. Excellent communications can be accomplished even though the response of the audio amplifiers used in military equipment seldom extends over a wide range of audio frequencies.

The output from the audio amplifier is then sent to a reproducer. The reproducer in voice communications may be either earphones or a loudspeaker. In either case, the function of the reproducing device is to convert the electrical signal variations into mechanical or audible variations. The speakers or phones used to accomplish this function are conventional.

## 49-15. Antenna Relay

The antennas chosen for use with the UHF transceiver are those which can handle a wide range of frequencies with almost complete absence of reflected waves in either the antenna or the associated transmission line. These antennas may be directional or non-directional as the application may dictate. In most Naval applications, a omni-directional antenna would be preferred for shipboard uses because it is not likely that one station will know the exact location of the other station. One of the types of antennas which will satisfy this requirement is the simple dipole antenna.

The function of the UHF antenna is to radiate or receive electromagnetic energy. The antennas chosen for UHF transceivers have the ability to transmit as well as receive. The device that permits one antenna to be used for both transmitting and receiving is called the antenna relay. The antenna relay transfers the same antenna from the receiver to the transmitter when the current through its exciting coil is started or stopped. The component parts of a relay are a coil wound on an iron core and an armature that operates a set of contacts. A simple relay and electrical connections to the transceiver are shown in Figure 49-23.

Closing the mike button allows current to flow through the coil energizing the electromagnet and drawing the armature downward, thereby, transfering the antenna from the receiver to the transmitter. The contacts through

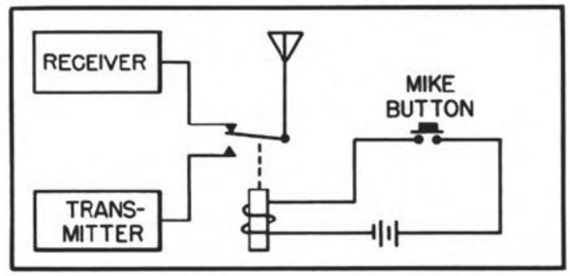


Figure 49-23 - Simple relay circuit.

which the RF current flows are usually made of special low resistance material. More contacts can be added to the armature so that other functions may be accomplished at the same time.

Upon releasing the mike button, the magnetic field about the relay coil collapses and the armature is returned to the receiver position by a mechanical spring arrangement. This spring arrangement and multiple contacts are illustrated in Figure 49-24.

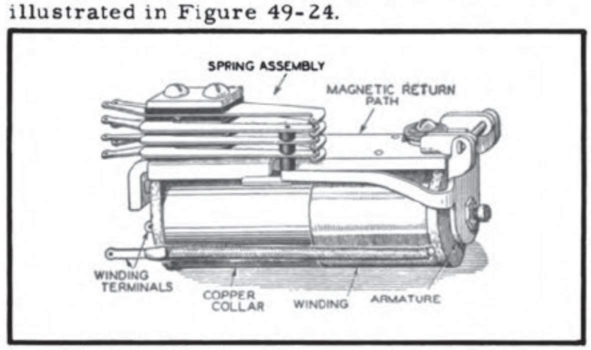


Figure 49-24 - Common type of relay.

Antenna relay coils may be designed to operate on either ac or dc. AC relays must be made rather heavy to prevent chatter and vibrations caused by the alternations of the magnetic field.

## EXERCISE 49

1. Describe the difference between the surface wave and the sky wave.
2. Is it possible to receive a UHF signal even though the transmitting antenna can not be seen from the receiving antenna? Explain.
3. Why are the tank circuits of a TPTG oscillator tuned to operate below their natural resonant frequency?
4. Explain why the value of capacitance for $C_{gp}$ in the TPTG oscillator should be relatively small.
5. Briefly explain the procedure for tuning a TPTG oscillator.
6. Explain the term piezoelectric effect.
7. What is the primary result of non-uniform thickness in a crystal?
8. Which crystal cut exhibits the most desirable temperature coefficient for frequency stability?
9. Explain why the high effective Q of the crystal aids in determining the accuracy of oscillations.
10. Why is a frequency multiplier necessary in a UHF receiver or transmitter?
11. Describe the operation of a push-pull tripler.
12. What is a tuned line? Describe how a tuned line will exhibit different characteristics at various distances from the source.
13. Describe a method of varying the characteristics (tuning) of a tuned line.
14. List and explain the two main limitations imposed by tube structure on the operation of a conventional electron tube at UHF frequencies.
15. Explain why the Q of a tuned line can be made much higher than a conventional LC tank at UHF frequencies.
16. What is transit time? How does it affect the operation of an amplifier?
17. Describe the different types of noises generated in vacuum tubes.
18. Describe the operation and advantages of the grounded-grid amplifier.
19. What is the difference between a crystal oscillator and a crystal controlled oscillator?
20. What is double conversion? Why is it used? What is its advantages?

# CHAPTER 50

# PRINCIPLES OF RADAR

## ELEMENTS OF RADAR

The word RADAR is formed as an abbreviation for RAdio Detection And Ranging. RADAR is an electronic device that may be used to detect the presence of objects such as airplanes or ships even in darkness, fog, or storm. In addition to indicating their presence, radar may be used to determine their bearing and distance. In special types, elevation and speed may also be indicated. It is one of the greatest scientific developments that has emerged from World War II. Its development, like that of most great inventions, was mothered by necessity—that of detecting the enemy before he detected us. The basic principles on which its functioning depends are relatively simple, and the seemingly complicated series of electrical events encountered in radar can be resolved into logical series of functions, which, taken individually, may be identified and understood.

## PRINCIPLES OF OPERATION

The principle upon which radar operates is very similar to the principles of sound echoes or wave reflection.

### 50-1. Sound Wave Reflection

If a person shouts in the direction of a cliff, or some other sound-reflecting surface, he hears his shout "return" from the direction of the cliff. What actually takes place is that the sound waves, generated by the shout, travel through the air until they strike the cliff. There they are reflected or "bounced off," and some are returned to the originating spot, where the person is then able to hear the echo. Some time elapses between the instant the sound originates and the time when the echo is heard because sound waves travel through air at approximately 1100 feet per second. The farther the person is from the cliff, the longer this time interval will be. If a person is 2200 feet from the cliff when he shouts, about 4 seconds elapses before he hears the echo—that is, 2 seconds for the sound waves to reach the cliff and 2 seconds for them to return.

If a directional device is built to transmit and receive sound, the principles of echo together with a knowledge of the velocity of sound can be used to determine the direction, distance, and height of the cliff shown in Figure 50-1. A sound transmitter, which can generate pulses of sound energy, can be so placed at the focus of the reflector that it radiates a beam of sound. The sound receiver can be a highly directional microphone located inside a reflector (at its focal point, and facing the reflector) to increase the directional effect. The microphone is connected through an amplifier to a loudspeaker.

Then, to determine the distance and direction of the cliff the transmitting and receiving apparatus are placed so that the line of travel of the transmitted sound beam and the received echo will very nearly coincide. They would coincide exactly if the same reflector could be used for both transmitting and receiving, as is done in radar systems. The apparatus (both the transmitter and receiver) is rotated until the maximum volume of echo is obtained. The horizontal distance to the cliff can then be computed by multiplying one-half of the elapsed time in seconds by the velocity of sound. This will be essentially the distance along the line RA (Figure 50-1, A). If the receiver has a circular scale that is marked off in degrees, and if it has been properly orientated with a compass, the direction or azimuth of the cliff can be found. Thus, if the angle indicated on the scale is $45^{\circ}$, the cliff is northeast from the receiver position.

To determine height (Figure 50-1, B), the transmitter and receiver antennas are tilted from the horizontal position (shown by dotted lines) while still pointing in the same direction. At first the echo is still heard, but the elapsed time is increased slightly. As the angle of elevation is increased, an angle is found where the echo disappears. This is the angle at which the sound is passing over the top of the cliff and is therefore not reflected back to the receiver. The angle at which the echo just disappears is such that the apparatus is pointing along line RB. If the receiver is equipped with a scale that permits a determination of the angle of elevation, the height of the cliff, AB, can be calculated from this angle and either the distance RA or RB, by the use of one of the basic trigonometric ratios.

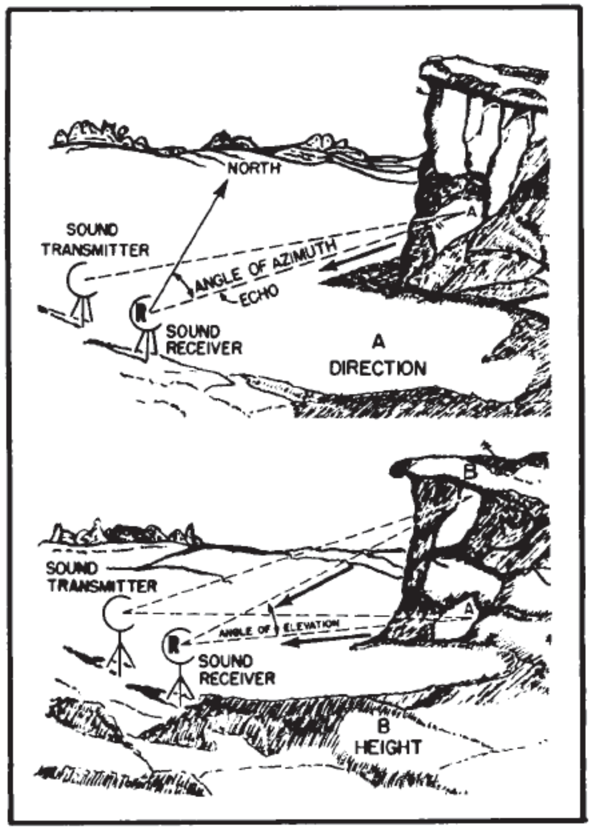


Figure 50-1 - Determination of direction and height.

Q1. In reference to Figure 50-1B, if the angle of elevation is $31^{\circ}$, and distance RA is 40 feet, what is the approximate height of the cliff?

## 50-2. Radio-Wave Reflection

All radar sets work on a principle very much like that described for sound waves. In radar sets, however, a radio wave of extremely high frequency is used instead of a sound wave. The energy sent out by a radar station (Figure 50-2) is similar to that sent out by an ordinary radio transmitter.

The radar station has one outstanding characteristic different from a radio in that it picks up its own signals. It transmits a short pulse and receives its echoes, then tranmits another pulse and receives those echoes. This out-and-back cycle is repeated 60 to 4000 times per second, depending on the design of the set. If the outgoing wave is sent into clear space, no energy is reflected back to the receiver. The wave, and the energy that it carries, simply travels out into space and is lost for all practical purposes.

If, however, the wave strikes an object such as an airplane (Figure 50-2), a ship, a building, or a hill, some of the energy is sent back as a reflected wave. If the object is large compared to a quarter-wavelength of the transmitted energy, a strong echo (but only a fraction of the transmitted energy) is returned to the antenna. If the object is small, the reflected energy is small and the echo is weak.

Radio waves travel at the speed of light, approximately 186,000 land miles per second or 162,000 nautical miles per second. A physical concept of this speed may be gained by considering the circumference of the earth as approximately 21,770 nautical miles. A radio wave could encircle the earth in approximately 134,400 $\mu$s (microseconds), or in slightly less than one-seventh of a second. Fifteen trips around the earth would be accomplished in slightly more than 2 seconds. This amounts to 2,000,000 $\mu$s. Consider that the moon is about 206,000 nautical miles (240,000 land miles) from the surface of the earth. Radar signals have been directed toward the moon and their echoes were returned. The elapsed time was approximately 2 1/2 seconds.

Most of the radio waves of the uhf and shf bands are only slightly affected by the earth's atmosphere and travel in straight lines. Accordingly, there will be an extremely short time interval between the sending of the pulse and the reception of its echo. It is possible, however, to measure the interval of elapsed time between the transmitted and received pulse with great accuracy—even to one ten-millionth of a second ($1 \times 10^{-7}$ seconds). The forming, timing, and presentation of these pulses are accomplished by a number of special circuits and devices.

The directional antennas employed by radar equipment transmit and receive the energy in a fairly sharply defined beam. Therefore, when a signal is picked up, the antenna can be rotated until the received signal is maximum. The direction of the target is then determined by the position of the antenna.

The echoes received by the radar receiver appear as marks of light on an oscilloscope (called "scope" for short). This scope may be marked with a scale of miles (or yards), or degrees, or both. Hence, from the position of a signal echo on the scope, an observer can tell the range and bearing of the corresponding target.

## 50-3. Historical Development

One of the first observations of "radio echoes" was made in the United States in 1922 by Dr. A. H. Taylor at the Naval Research Laboratory. Dr. Taylor observed that a ship passing between a radio transmitter and receiver reflected some of the waves back toward the transmitter. Between 1922 and 1930 further tests proved the military value of this principle for the detection

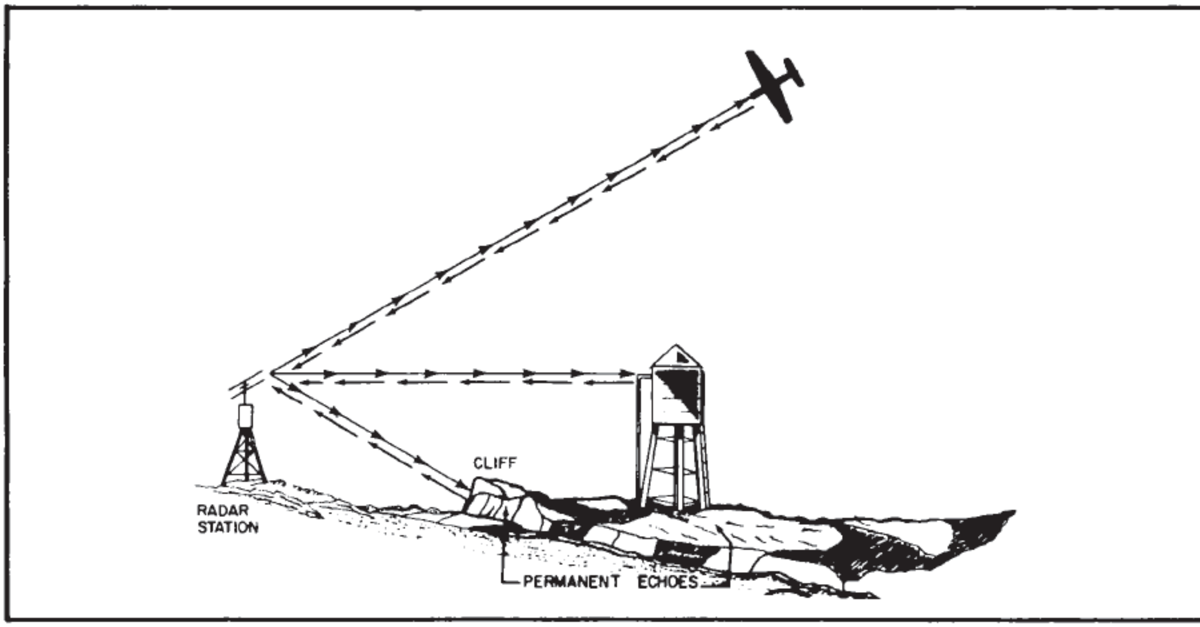


Figure 50-2 - Transmission and reflection of radar pulses.

of objects that would normally be hidden by smoke, fog, or darkness. During this same period Dr. Breit and Dr. Tuve of the Carnegie Institute published reports on the reflection of pulse transmissions from electrified layers in upper atmosphere. This led to the application of the principle to the detection of aircraft. Other countries carried on further experiments independently and with utmost secrecy. By 1936, the United States Army was engaged in the development of a radar warning system for coastal frontiers. By the end of 1940 the British had developed radar to such a point that they were able to bring down great numbers of enemy airplanes with guns being accurately controlled by radar systems. Beginning in 1941, British-American cooperation in the development of radar gave the Allies the best radar equipment in the world.

Along with the development of radar came the development of effective countermeasures. Since 1941 great advances have been made in radar and in countermeasures in the various research and development centers throughout the country.

Q2. What basic factors would determine the strength of a returned radar echo?

## RADAR DETECTING METHODS

### 50-4. Continuous-Wave Method

The continuous-wave (CW) method of detecting a target makes use of the Doppler effect. The frequency of a radar echo is changed when the object which reflects the echo is moving toward or away from the radar transmitter. This change in frequency is known as the DOPPLER EFFECT. A similar effect at audible frequencies is recognized readily when the sound from the whistle of an approaching train appears (to the ear) to increase in pitch. The opposite effect (a decrease in pitch) occurs when the train is moving away from the listener. The radar application of this effect permits a measurement of the difference in frequency between the transmitted and reflected energy and thus a determination of both the presence and speed of the moving target. This method works well with fast-moving targets, but not well with those that are slow or stationary. CW systems are therefore limited in present usuage.

### 50-5. Frequency-Modulation Method

In the frequency-modulation (FM) method the transmitted energy is varied continuously and periodically over a specified band of frequencies. The instantaneous frequency of the energy being radiated by the antenna therefore differs from the instantaneous frequency being received by the antenna.

The frequency difference depends on the distance traveled and can be used as a measure of range. Moving targets produce a frequency shift in the returned signal because of the Doppler effect, however, and this affects the accuracy of range measurements. This method, therefore, works better with stationary or slow-moving targets than with fast-moving ones.

A1. $\tan\theta = \frac{opp}{adj}$; $\tan 31^\circ = \frac{opp}{40}$ = opp = .6(40) = 24 feet.

A2. Strength of transmitted pulse, size of target, and distance of target.

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### 50-6. Pulse-Modulation Method

In the pulse modulation method the RF energy is transmitted in short pulses in which the time duration may vary from 0.1 to 50 $\mu$s. If the transmitter is turned off before the reflected energy returns from the target the receiver can distinguish between the transmitted pulse and the reflected pulse. After all reflections have returned, the transmitter can be turned on again and the process repeated. The receiver output is applied to an indicator that measures the time interval between the transmission of the energy and its return as a reflection. Because the energy travels at a constant velocity, one-half the time interval between the outgoing pulse and its echo becomes a measure of the distance traveled by the pulse to the target, or the range. Because this method does not depend on the relative frequencies of the emitted and returned signals or on the motion of the target, difficulties experienced in the CW and FM methods are not present. The pulse-modulation method is used almost completely in military applications. Therefore, it will be the only method discussed in this text.

Q3. What three factors, using pulse-modulation, must be known to determine the position of an object in space?

## USES OF RADAR

Naval scientists pioneered in finding practical uses for radar. Those uses were made chiefly for detecting and destroying an enemy and his armaments; this is still the most important naval use today. Civilian uses followed those made for military purposes. For example, some familiar civilian uses are (1) radar speed determination on highways for controlling traffic, (2) radar weather prediction, (3) commercial radar air navigation, and (4) safeguarding air vehicles and merchant ships from collision hazards.

Since no single radar appliance can yield all of the information required of naval shipboard systems, several classes, or types, have been designed. Each of these types, or categories, has its limitations and capabilities within the purpose for which built. Naval shipboard radar equipments are grouped in three general categories: search, fire-control, and special.

Search radars are of two categories: air search and surface search. These equipments are used for early warning networks and for general navigational purposes. Search radars produce detection at maximum ranges while sacrificing some degree of accuracy and resolution (detail).

Fire-control radars, integral parts of certain gunfire control systems, are used after targets have been located by search radars.

Special radars are used for specific purposes, which include recognition, or identification of friend or foe (IFF), ground-controlled and carrier-controlled approach (GCA and CCA, respectively), range rate or speed, and height finding.

### 50-7. Search Radar

Search radars used for early warning nets do not require great precision in ranging or bearing, but do require the ability to locate targets at fairly long ranges. Therefore, they are normally designed with high power, wide beam angle, and fairly long pulse widths. Their target resolution (ability to accurately determine bearing and range) is not as good as that of radars used for another purpose such as fire control.

Each type of radar equipment has been designed for definite purposes. An air search radar performs inadequately for surface search and vice versa. Each of these types may be used in an emergency as fire control radar, but cannot be expected to furnish bearings, ranges and position angles with the same degree of accuracy as radar designed for that purpose.

### 50-8. Air Search Radar

Primary function of an air search radar is the detection and determination of ranges and bearings of aircraft targets at long ranges maintaining complete 360° surveillance from the surface to high altitudes. System constants must be chosen with this function in mind. Relatively low radar frequencies are chosen (P or L band 300-1000 mc) to permit long-range transmissions with minimum attenuation. Wide pulse widths (2 to 4 $\mu$s) and high peak power are used to aid in detecting small targets at great distances. Low pulse repetition rates are selected to permit greater maximum measurable range. Wide vertical beam width is used to ensure detection of targets from the surface to relatively high altitudes, and to compensate for the pitch and roll of the ship. Medium horizontal beam width is employed to permit fairly accurate bearing resolution while maintaining 360° search coverage.

## 50-9. Surface Search Radar

The primary function of surface search radar is the detection and determination of accurate range and bearing of surface targets while maintaining $360^{\circ}$ surveillance for all surface targets within line-of-sight distance of the radar antenna.

Since the maximum range requirement of a surface search radar is primarily limited by the radar horizon, very high frequencies (X band) are employed to permit maximum reflection from small target-reflecting areas, such as ship mast-head structures and submarine periscopes. Narrow pulse widths (0.37 to 2 $\mu s$) are used to permit a high degree of range resolution at short ranges, and to achieve greater range accuracy. High-pulse repetition rates (600 to 1000) are used to permit maximum illumination of targets. Medium peak powers can be used to permit detection of small targets at line-of-sight distances. Wide vertical beam widths ($10^{\circ}$ to $30^{\circ}$) permit compensation for pitch and roll of own ship and to detect low-flying aircraft. Narrow horizontal beam widths ($1^{\circ}$ to $3^{\circ}$) permit accurate bearing determination and good bearing resolution.

## 50-10. Fire Control Radar

The primary function of fire control radar is the acquisition of targets originally detected and designated from search radars, and the determination of extremely accurate ranges, bearings, and position angles of targets within firing range. The antennas can be stabilized to compensate for pitch and roll of own ship. Very high frequencies are chosen to permit the formation of narrow beam widths with comparatively small antenna arrays, detection of targets with small reflecting areas, and high detail of all targets. Very narrow pulse widths provide a high degree of range accuracy, at short range, and excellent range detail. Very high repetition rates afford maximum target illumination while using very narrow pulse widths. Since long ranges are not required, low peak power permits the use of smaller components by keeping the average power low. Narrow, vertical, and horizontal beam widths provide accurate bearing and position angles and a high degree of bearing and elevation resolution.

## 50-11. Airborne Radar

Radar equipments for aircraft are of the same general types as for land or shipboard except that they are physically much smaller. Both search and fire control radars are successfully used in aircraft. While radar is a powerful aid to aircraft, the aircraft in turn increases the range of radar by supplying it with an elevated platform from which its effective range in detecting objects is greatly extended because the line-of-sight distance is increased toward a farther horizon.

Radar information picked up in a plane may be relayed by radio transmitter to another distant location on board ship or elsewhere, thereby effectively increasing the range.

## 50-12. Auxiliary Equipment

Some means of distinguishing a friendly target from an enemy target is necessary. An electronic system for "identification, friend or foe" (IFF system) is used for this purpose. This auxiliary equipment provides an accurate and rapid means of determining the friendly or enemy character of objects detected by radar.

The IFF system equipment consists of two groups containing two units in each group. Two units (transmitter-receiver) are located with the radar. The other two (receiver-transmitter) are located in friendly craft. The transmitter-receiver group is referred to as the RECOGNITION SET (also INTERROGATOR) and the receiver-transmitter group is called the IDENTIFICATION SET (also TRANSPONDER)

When a radar operator observes an unidentified target on his radar, he sets the first group (the INTERROGATOR) in operation. The interrogator transmitter is a pulse-type transmitter, which emits coded challenging pulses. The transpondor's receiver receives these pulses, which trigger (automatically) a coded reply known only by the friendly operators. The transpondor's transmitter releases this coded reply, which is received by the interrogator's receiver and placed on an indicator for evaluation. The indicator may be either an integral section within the radar scope or a separate scope.

## 50-13. Information Produced by Radar

Radar increases the effectiveness of naval craft by adding new powers and capabilities to the human senses. It is unhampered by the ordinary obstacles to unaided vision such as darkness, fog, haze, and smoke. Radar reveals the presence and location of certain kinds of objects situated far beyong the range of normal vision, indicating their distance and bearing directly and with a high degree of accuracy. Radar pierces the surrounding darkness or overcast and reveals aircraft, ships, land areas, cities, cloud, and hazards to navigation.

Q4. Why would low radar frequencies be chosen?

Q5. How does an IFF system differ from a search radar?

A3. The range, azimuth and elevation.

A4. To permit long-range transmissions with minimum attenuation.

A5. The IFF system relies on a properly transmitted signal and not an echo.

## TYPES OF PRESENTATION

To furnish usuable information, a radar set must produce some type of visual presentation of the target echo so as to suggest a mental image of the target to the observer. Cathode-ray tubes are used for this purpose. The scope images contain data of measurable quantities, including range, time, height, speed, and azimuth. Several types of data presentation have been developed to give the required information.

A radar beam systematically reveals what it SCANS (scrutinizes or examines in great detail). The results of each scan are revealed (presented) as a picture or presentation on the scope. About 15 types of scans have been designed, but naval requirements can be met by use of only a few of them as can be seen in Figure 50-3. Each type of scan is identified by a letter of the alphabet.

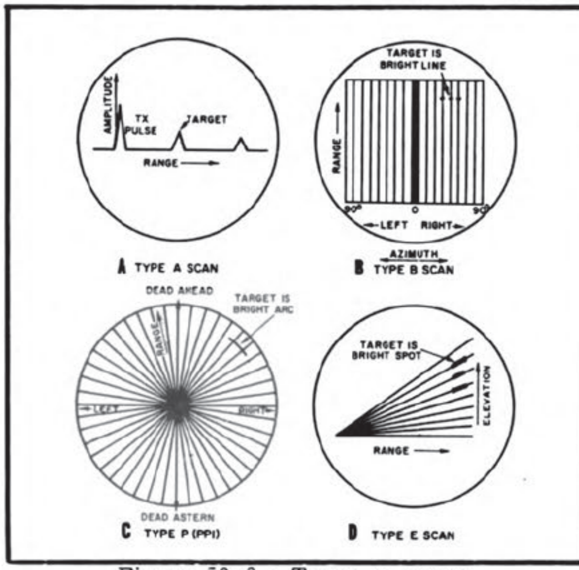


Figure 50-3 - Types of scans.

### 50-14. Type-A Scan

Type-A presentation is used to determine range. The screen of this scope has a short persistence. The echo causes a vertical displacement of the spot, the amplitude of which depends on the strength of the returned signal pulse. The point on the horizontal base line at which the vertical displacement occurs indicates the range. Type-A presentation is shown in Figure 50-3, A.

### 50-15. Type-B Scan

The type-B presentation (Figure 50-3, B) indicates both range and azimuth angle (bearing). The vertical displacement of the echo signal indicates range, and the horizontal displacement of the echo signal indicates azimuth angle. This scope has long persistence.

### 50-16. Type-PPI Scan

The PPI (Plan Position Indicator) presentation is another type of scan for presenting range and bearing (azimuth) information. See Figure 50-3, C. You can think of the PPI scan as a modified type-B scan, in which rectangular coordinates are replaced by polar coordinates.

The antenna is rotated uniformly about the vertical axis so that searching is accomplished in a horizontal plane. The radar beam is usually narrow in azimuth and broad in elevation. Large numbers of pulses are transmitted for each rotation of the antenna. As each pulse is transmitted, the spot starts from the center of the indicator and moves toward the edge along a radial line. Upon reaching the edge of the scope, the spot quickly jumps back to the center and begins another trace as soon as the next pulse is transmitted. As the antenna rotates, the path of the spot also rotates around the center of the scope so that the angle of the radial line on which the spot appears indicates the azimuth of the antenna beam, and distance (out from the center of the scope) indicates the range.

When an echo is received, the intensity of the spot is increased considerably, and a brighter spot remains at that point on the screen, even after the scanning spot has passed it. Thus, it is possible with this scan to produce a map of the territory surrounding the observing station on the scope. This type of scan is useful when the radar set is used as an aid to navigation.

In type-A presentation the echo signal causes vertical deflection of the trace—in other words, it is DEFLECTION MODULATED. In type-B and type-PPI presentation the echo signal makes the trace brighter. This action is called INTENSITY MODULATION.

### 50-17. Type-E Scan (RHI).

The RHI (Range Height Indicator) presentation is another type of scan for presenting range and height information. The RHI scan is also known as the type-E scan shown in Figure 50-3, D. The type-E scan is a modification of the type-B scan on which an echo appears as a bright spot with

the range indicated by the horizontal coordinate and the elevation (height) as the vertical coordinate. This type is used (1) in directing planes in blind landing, (2) for ground-controlled approach, (3) for carrier-controlled approach, and (4) in determining altitude. This scan is also used in the Mk 56 gunfire control system (GFCS), which employs the Mk 35 radar. There the presentation is specified as the E (delta E) scan, which means that the elevation changes are presented.

Q6. Of the various type scans, what is the difference between intensity modulation and deflection modulation?

Q7. Which of the various type scans employ the horizontal axis to indicate range information?

Q8. Which of the various type scans shows the elevation of a target?

## RANGE AND BEARING

### 50-18. Range Determination

The successful employment of pulse-modulated radar systems depends primarily on the ability to measure distance in terms of time and knowledge of the velocity of light. Radio-frequency energy, once it has been radiated into space, continues to travel with a constant velocity. When it strikes a reflecting object there is no loss in time, but merely a redirecting of the energy. Its velocity is that of light, or, in terms of distance traveled per unit of time, 186,000 landmiles per second, 162,000 nautical miles per second, or 328 yards per microsecond. This means that it takes approximately 6.1 μs for radio energy to travel 1 nautical mile, or approximately 6080 feet (2027 yards). All radar ranging is based on a flat figure of 6080 feet per mile and, because the speed of light (and radio waves) is so great, microseconds (μs) are used for all time determination.

This constant velocity of radio-frequency energy is applied in radar to determine range by measuring the time required for a pulse to travel to a target and return. The time lapse between the transmitted pulse and the echo return may be readily determined with the aid of the oscilloscope. For the purpose of illustrating how this may be done, assume that a target ship is 20 nautical miles away from the radar transmitter-receiver combination. Because radio energy travels 1 nautical mile in 6.18 microseconds, 123.6 microseconds will be required for the transmitted pulse to reach the target, or a total of 247.2 microseconds before the echo will return to the radar receiver.

The horizontal sweep frequency of the scope is adjusted so that it makes one complete sweep (from left to right) during the time the transmitted pulse is going to the target (maximum range) AND THE ECHO IS RETURNING TO THE RECEIVER. In other words, the time of one sweep is 247.2 microseconds, and the frequency is therefore approximately 4045 cps. Assume that a translucent scale with uniform divisions in miles from 0 to 20 is placed over the face of the scope; and assume further that the extent of the sweep extends from the 0 mark to the 20-mile mark. In this case the maximum range is 20 miles.

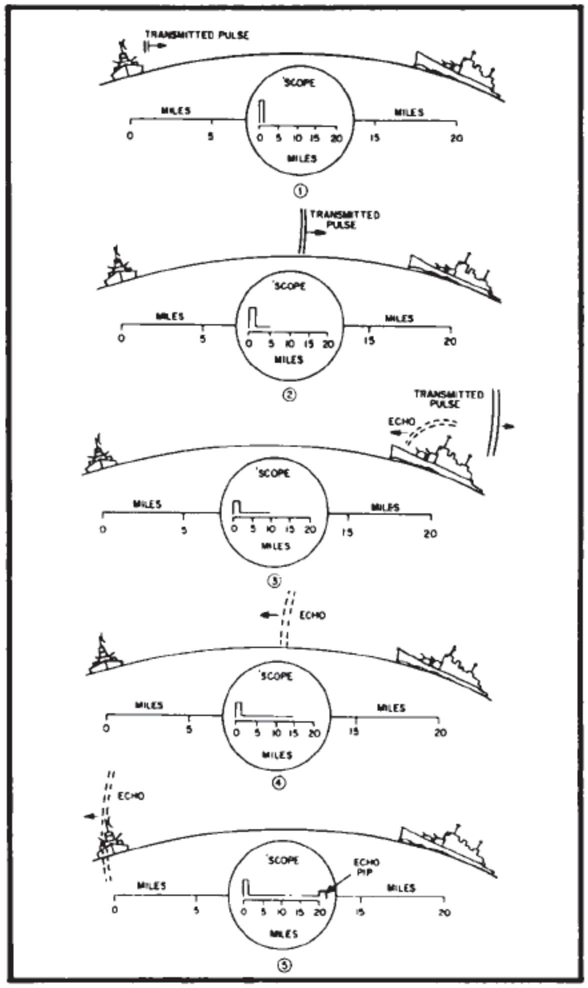


Figure 50-4 - Radar range determination.

Figure 50-4 shows how the range to the target is determined. (1) the transmitted pulse is just leaving the antenna. A part of the generated energy is fed to the vertical deflection plates at the instant the pulse is transmitted and causes a vertical line (pip) toa ppear at the zero-mile mark on the scope. (2) 61.8 microseconds later, the transmitted pulse has traveled 10

A6. In deflection modulation, the echo signal causes a vertical variation of the trace (indicator presentation); in intensity modulation the echo signal causes the trace to increase in brightness.

A7. The A and E scans.

A8. The E type scan.

miles toward the target. The horizontal trace on the scope, however, has reached only the 5-mile mark—that is, one-half the distance the transmitted pulse has traveled (the sweep frequency is timed to indicate one-half the distance. (3) 123.6 microseconds after the initial pulse left the transmitter, the transmitted pulse has reached the target, 20 miles away and the echo has started back. The scope reading is 10 miles. (4) 185.4 microseconds after the initial pulse, the echo has returned half the distance from the target, and the scope reading is 15 miles. (5) 247.2 microseconds after the initial pulse, the echo has returned to the receiving antenna. This relatively small amount of energy is amplified and applied to the vertical deflection plates, and an echo pip of smaller amplitude than the initial pip is displayed on the scope at the 20-mile mark.

If two or more targets are in the path of the transmitted pulse each will return a portion of the incident energy as echoes. The targets farthest away (assuming they are similar in size and type of material) will return the weakest echo.

In conjunction with the scope there is a hand-crank and mechanical counter assembly that enable the operator to determine the range with a greater degree of accuracy. When a target is indicated on the base line the operator turns the handcrank to move the range indicator, or gate (Figure 50-5), to the target and then reads the range, in yards, directly from the counter assembly. This process is known as "gating the target."

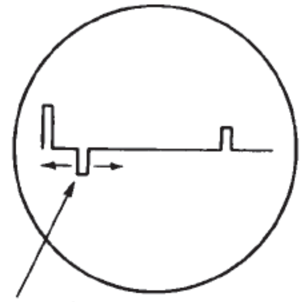


Figure 50-5 - Target gating.

Q9. What does target gating accomplish?

50-19. Bearing Determination.

The bearing (true or relative) of the target may be determined if the direction in which the directional antenna is pointing when the target is picked up is known. Control and indicator systems have been devised that make this possible.

The measurement of the bearing of a target as "seen" by the radar is usually given as an angular position. The angle may be measured either from true north (true bearing), or with respect to the heading of a vessel or aircraft containing the radar set (relative bearing). The angle at which the echo signal returns is measured by utilizing the directional characteristics of the radar antenna system. Radar antennas are constructed of radiating elements, reflectors, and directors to produce a single narrow beam of energy in one direction. The pattern produced in this manner permits the beaming of maximum energy in a desired direction. The transmitting pattern of an antenna system is also its receiving pattern. An antenna can therefore be used to transmit energy, to receive reflected energy, or to do both.

The simplest form of antenna for measuring azimuth or bearing is one that produces a single-lobe pattern. The system is mounted so that it can be rotated. Energy is directed across the region to be searched, by moving the beam back and forth in azimuth until a return signal is picked up. The position of the antenna is then adjusted to give maximum return signal.

Figure 50-6 shows the receiving pattern for a typical radar antenna. In this figure, relative signal strength is plotted against the angular position of the antenna with respect to the target. A maximum signal is received only when the axis of the lobe passes through the target. The sensitivity of this system depends on the angular width of the lobe pattern. The operator adjust the position of the antenna system for maximum received signal. If the signal strength changes appreciably when the antenna is rotated through a small angle, the accuracy with which the on-target position can be selected is great. Thus, in Figure 50-6, the relative signal strengths A and B have very little difference. If the energy is concentrated in a more narrow beam, the difference is greater and the accuracy better.

Q10. What is the advantage of a narrow trans-

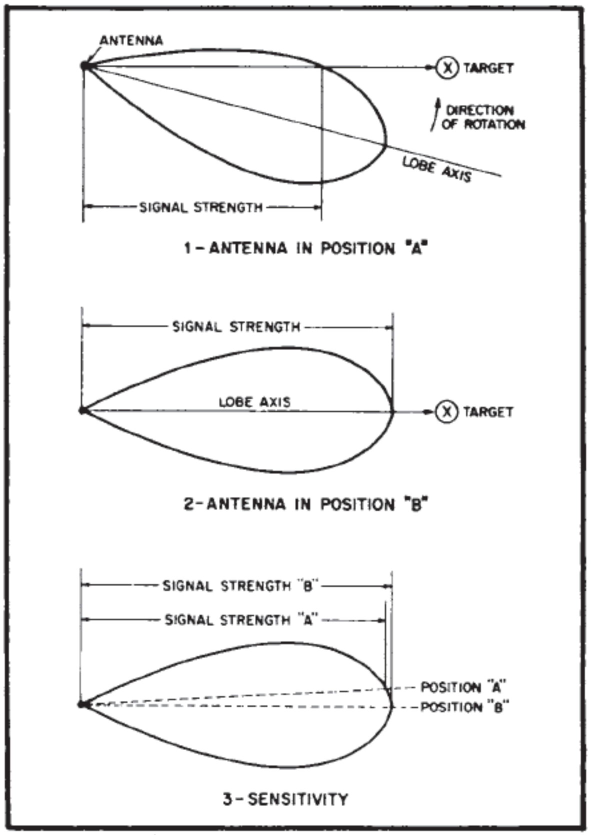


Figure 50-6 - Radar determination of aximuth or bearing.

mitted radiation signal pattern when determining aximuth location of a target?

50-20. Plan Position Indicator

The range scope has certain limitations when it is desired to know what is happening instantaneously in all directions because it indicated only the targets in the direction in which the antenna is instantaneously pointing.

A master PPI allows the radar operator to see the screen images of all objects surrounding his craft (within the range limitations of the equipments because it displays a graphic plot of 360° of antenna rotation and has a screen of the necessary persistence to retain the targets visible after the antenna has rotated past the target bearing.

The range scope presents the target information on a horizontal base line as shown in Figure 50-3, A. The PPI has a radial base line originating at the center of the screen (Figure 50-3, C) which indicates the physical antenna location, and this line follows the antenna rotation.

A view of the PPI scope is shown in Figure 50-7. The bright spots on the screen are images of objects (ships, planes, land masses, etc.) in the vicinity of the craft carrying the PPI

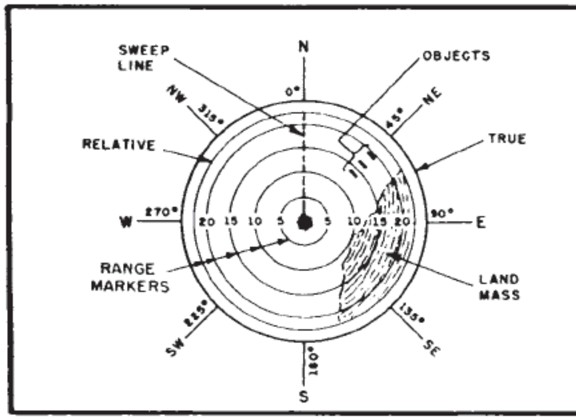


Figure 50-7 - PPI presentation.

equipment. Around the outer edge of the scope are relative and true bearing circles. Spaced evenly across the face of the tube are range rings, calibrated in miles. Thus, from the position of the images, their approximate range and bearing may be determined from the scope. A particular object of interest may be singled out for more accurate ranging by referring to the range scope.

Another principle difference between the two systems (range and PPI) is the method of applying the signal to the scope. In the RANGE scope the echo signal is amplified and applied to the vertical deflection plates in such a way as to produce a pip on the horizontal time-base line, on the screen. In the PPI SCOPE, the echo signal is amplified and applied to the control grid or cathode of the scope in such a way that the trace is brightened momentarily on the radian time-base line. If the intensity of the trace is kept sufficiently low, the scope will be essentially dark until an echo is received, and then the contrast will be very pronounced.

The PPI uses electromagnetic deflection instead of electrostatic deflection. Current flowing from the sweep generator through a single pair of electromagnets mounted across the neck of the tube at right angles to the axis of the tube causes the electron beam to be swept from the center of the tube to one edge and back again to the center.

In earlier types of PPIs the deflection electromagnets were mounted so that they can be rotated around the neck of the tube. The rotating assembly is synchronized with the antenna rotation so that when the antenna turns, the sweep trace is rotated about the screen at the same rate.

Thus, for example, in Figure 50-7 when the antenna is pointing in the NE direction the deflection magnets will force the beam across the screen from the center to the outer edge in the NE direction. The beam will be deflected across the screen many times during the course of a

A9. Enables the operator to determine the range of a target with a greater degree of accuracy.

A10. The more narrow the pattern, the more accurately the position of the antenna system may be adjusted for maximum received signal strength.

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small angular rotation of the magnets. In this area on the screen the echoes from the three targets will cause three areas of intensification on the screen.

Q11. What allows the target echo to remain on the scope screen after the sweep has passed it?

### 50-21. Altitude Determination

The remaining dimension necessary to locate completely an object in space can be expressed either as an angle of elevation or as an altitude. If one is known, the other can be calculated from one of the basic trigonometric ratios. A method of determining the angle of elevation or the altitude is shown in Figure 50-8. The slant range (Figure 50-8, A) is obtained from the radar scope indication as the range to the target. The angle of elevation is that of the radar antenna (Figure 50-8, B). The altitude is equal to the slant range multiplied by the sine of the angle of elevation.

In radar equipments with antennas that may be elevated, altitude determination by slant range is automatically computed electronically.

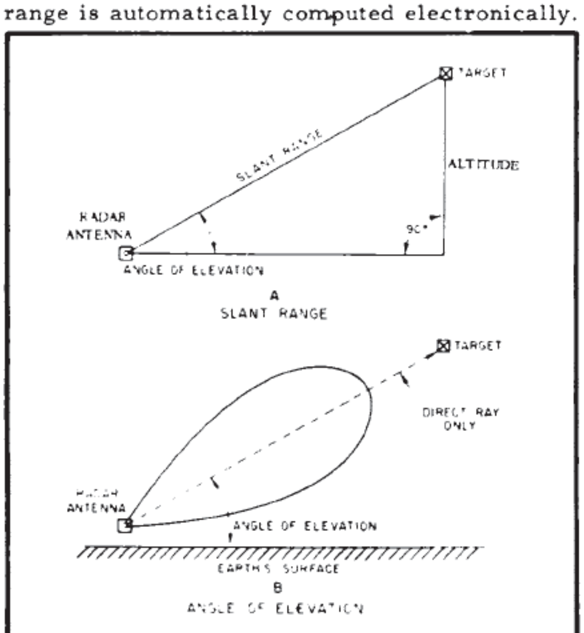


Figure 50-8 - Radar determination of altitude.

In equipments (air search) where the antennas do not elevate, the altitude may be calculated by means of "fade charts." A method for producing the chart and its application are described in the Navy training course for RADARMAN 3&2, NavPers 10146.

## FUNCTIONAL BLOCK DIAGRAM

Radar systems now in existence vary greatly in detail. They may be very simple; or, if more accurate data are required, they may be highly refined. The principles of operation, however, are essentially the same for all systems. Thus a single basic radar system can be visualized in which the functional requirements hold equally well for all specific equipments.

In general, the degree of refinement of radar circuits increases with the frequency. The microwave region lends itself to a higher degree of precision in angular measurement, and for this reason modern radars operate at superhigh frequencies.

The functional breakdown of a pulse-modulated radar system generally includes six major components, as shown in the block diagram of Figure 50-9. The components may be summarized as follows:

1. The modulator produces the synchronizing signals that trigger the transmitter the required number of times per second. It also triggers the indicator sweep and coordinates the other associated circuits. In some sets an external trigger generator is used to synchronize all triggered units.
2. The transmitter generates the RF energy in the form of short, powerful pulses.
3. The antenna system takes the RF energy from

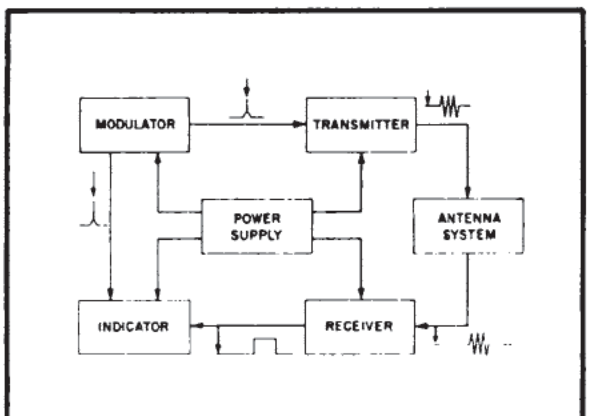


Figure 50-9 - Functional diagram of a fundamental pulse-modulated radar system.

the transmitter, radiates it in a highly directional beam, receives any returning echoes, and passes these echoes to the receiver.

4. The receiver amplifies the weak RF pulses returned by the target and reproduces them as video pulses to be applied to the indicator.
5. The indicator produces a visual indication of the echo pulses in a manner that furnishes the required information.
6. The power supply furnishes all ac and dc voltages necessary for the operation of the system components.

## 50-22. Radar Modulator

The function of the modulator is to ensure that all circuits connected with the radar system operate in a definite time relationship with each other and that the interval between pulses is of the proper length. In general, there are two practical methods of supplying the timing requirements—timing by means of a separate unit and timing within the transmitter.

A separate timing source may be used to give rigid control of the pulse-repetition-frequency. In this case the source consists of any stable type of audio oscillator such as the Wienbridge oscillator. The output is then applied to the necessary pulse-shaping circuits to produce the required timing pulse. Figure 50-10 shows in block form the functional components associated with the timer. These include the oscillator and other stages and components that are necessary to generate, shape, and amplify the waveform so that it may properly trigger the magnetron in the transmitter.

The oscillator generates a steady output at a given frequency (usually any frequency between 625 and 650 cps and generally less than 1000 cps), and this output establishes the PRF of the set.

The sine wave output of the oscillator is of the correct frequency but it does not have the correct shape and its amplitude is insufficient to fire the magnetron. Therefore, the signal is changed first into a square wave in the over-driven amplifier stage. The square wave is sharpened into a peaked wave in a differentiating circuit (a resistor and capacitor in series with the input, and the output taken across the resistor) and fed via a cathode follower to a blocking oscillator.

The blocking oscillator is triggered at the correct frequency by this peaked wave. The blocking oscillator generates the type of wave shape needed by the magnetron, except that it is of insufficient amplitude.

The output signal generated by the blocking oscillator is fed via a cathode follower to the power amplifier (preceded in actual circuits by driver amplifiers) where the square-wave pulse is amplified sufficiently to drive the magnetron. Only the negative portion of the pulse is used to drive the magnetron oscillator, and therefore the positive portion of the pulse is removed.

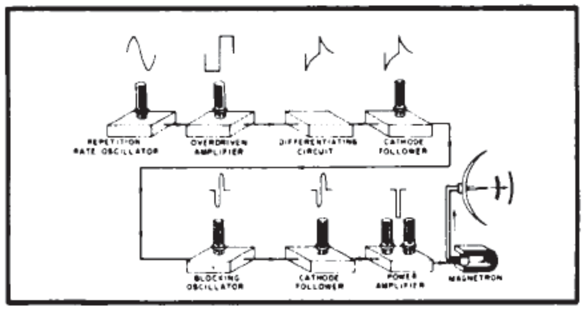


Figure 50-10 - Simplified block diagram of a modulator and transmitter.

The magnetron goes into oscillation the moment it is triggered by the negative-going square wave from the power amplifier. The frequency of the magnetron oscillation may be of the order of 6500 megacycles. The width of the pulse is determined by the width of the negative-going pulse from the power amplifier and may be of the order of 1 microsecond. During the pulse, the power output may be of the order of 125 kw.

Q12. To what blocks does the timer send a trigger?

Q13. What determines the frequency of the transmitted radar pulse?

## 50-23. Radar Transmitter

The transmitter is basically an RF oscillator. It may be turned on and off by the negative-going pulse from the modulator. The radar oscillator (in this instance a magnetron) differs from other oscillators treated in Chapter 24 in that it produces a much higher frequency and has a much higher power output. The higher frequency permits smaller waveguides and antennas to be used; and the higher power permits stronger echoes and a greater useful range.

Because of the superhigh frequencies in a radar set, buffers, frequency multipliers, and power output tubes following the magnetron would have little value in increasing the output power, and hence are not used in a radar set.

The more powerful sets are capable of putting out 1 megawatt (1000 kw) of peak power. A simplified diagram of a magnetron is shown in Figure 50-11,A. The magnetron is essentially a diode that has its plate at ground potential and its cathode at a high negative potential during the time it is oscillating. The diode is placed in a powerful magnetic field produced by a permanent magnet.

When a negative pulse is applied to the cathode and there is no magnetic field present, electrons move in straight lines from the cathode to the plate, as shown in (1) Figure 50-11, B. When

A11. The persistency of the phosphorescent coating on the CRT screen.

A12. The transmitter and the indicator sweep circuits.

A13. The magnetron.

a weak magnetic field (2) is applied, the electron paths become curved; and as the magnetic field becomes stronger ((2),(3),(4),and (5)), the electron paths become so curved that the electrons are moving in closed circular orbits that miss the plate entirely, and no plate current flows. The plate in Figure 50-11, C, is a copper cylinder the internal surface of which is separated into a number of segments by holes in the cylinder that serves as tuned circuits. As the electrons move in circles past the plate segments they induce currents electrostatically in the walls of the holes. The energy of the magnetron output pulse is contained in the field associated with these currents. **The frequency depends on the size of the cylinder, the strength of the magnetic field, and the difference in potential between the cathode and plate.**

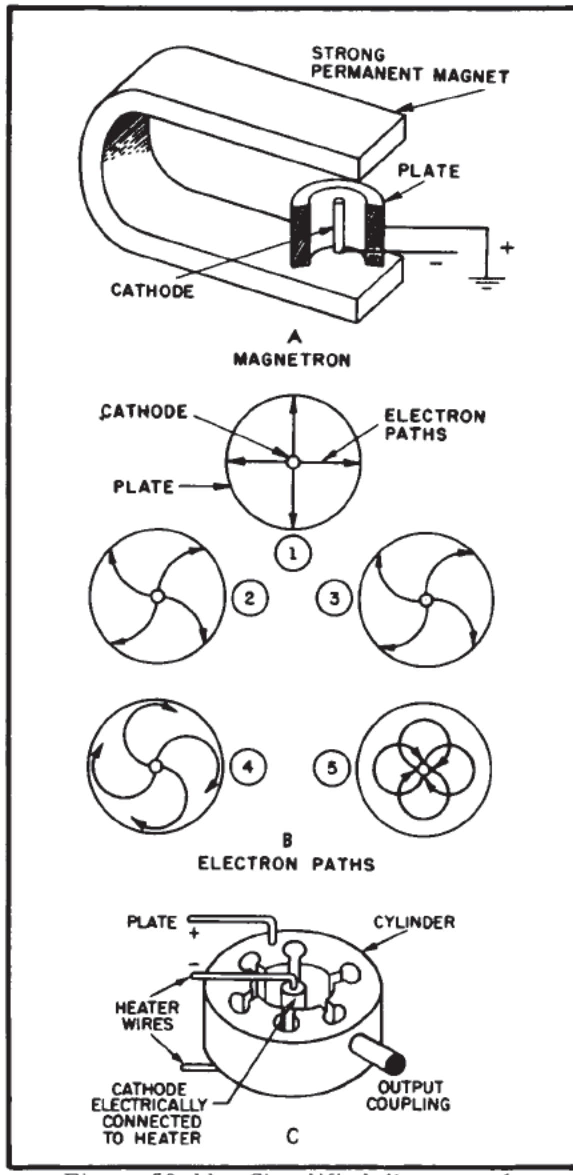

Figure 50-11 - Simplified diagram of a magnetron.

Energy is coupled out of the magnetron by means of a loop or probe; it is then transmitted to the antenna via a waveguide.

The tremendous peak power produced in short pulses by the magnetron requires high plate-to-cathode potential and high cathode emission. Because of the relatively long resting time between pulses, the problem of cooling is reduced and the physical size of the magnetron is not as large as would be expected from the peak power rating.

Q14. In the magnetron, which direction are the electrons traveling in relation to the magnetic field?

## 50-24. Radar Antennas

One function of an antenna system is to take the energy from the transmitter, radiate that energy in some chosen manner (by using a directional system when bearings are desired but by using a nondirectional antenna system where a bearing indication is not necessary). Another function of an antenna system is to pick up the returning echo, pass it to the receiver with a minimum of losses.

Some original radar installations contained two separate antenna systems: one for transmitting and one for receiving. The more practical radar system uses a single antenna system and an electronic switch capable of rapidly shifting the antenna performance from transmit to receive functions and vice versa. The switch is needed to protect the receiver from damage by the potent transmitter energy during the pulse time and, also, to keep the transmitter from absorbing some (or all) of the very weak echo during the receiving time.

Some applications of radar can use a simple nondirectional antenna, for example, the vertical dipole. Nondirectional antennas are used in navigation aids, as radar beacons (called racon), and some forms of IFF equipment.

That radar system which indicates the bear-

ings of targets must have some means of pointing its radiated energy in known directions. Practically all such radar systems accomplish this by constant 360° rotation of a motor-driven energy-transfer device such as an antenna, wave-guide, reflector or director, or energy feedhorn.

Each antenna type has abilities to couple and project electromagnetic energy into space; also each has an ability to convert received energy into the forms that activate receiver equipments.

Combining the characteristics from several antenna types results, obviously, in an improved system. Thus, a reflector may be added as shown in Figure 50-12, and driven by motor for continual 360° rotation, to direct concentrated energy toward the horizon. This provides a highly-directive antenna system for use at radar frequencies in vhf region and above. Figure 50-13 shows two antenna feed units at the focus; a dipole antenna for IFF purposes and flared feed-horn for search purposes. This dish is a section of a parabola.

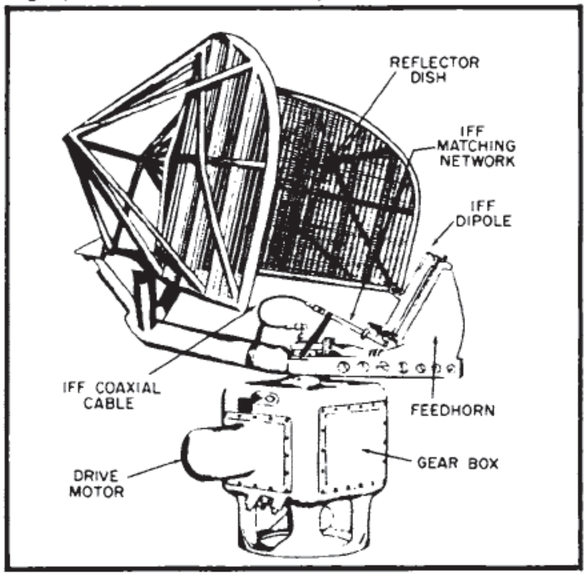


Figure 50-12 - Radar antenna with reflector.

Experience shows that a parabolic dish, when properly focused for projecting energy, will also serve at its best for accepting echo energy from space and returning it into the transmission system.

If the parabolic reflector is sufficiently large so the distance from any point within the dish to the focal point is several wavelengths, then QUASI-OPTICAL conditions will exist and the emerging wave is a narrow beam. Sizes of reflectors, which are practicable for microwave work, have a diameter of 10 to 20 wavelengths to produce a beam width of approximately 5 degrees.

The quasi-optical theory is mentioned many times in describing radar behavior. The word quasi means "similar" or "like." When you speak of microwaves from a high-frequency radar transmitter being quasi-optical waves in their behavior, you merely mean that invisible radar waves act like visible light waves.

## 50-25. Radar Receiver

When you compare Figure 50-14 with the equipment described in Chapter 31, you will recognize that the radar receiver is essentially a special type of superheterodyne receiver. Its function is to receive the weak echoes from the antenna system, combine them in a crystal mixer (half-wave crystal rectifier) with the RF signals from a local oscillator, amplify the resultant IF signal, detect the pulse envelope, amplify the resulting dc pulses, and feed them to the indicator. At the higher frequencies used in radar, it is not possible to use a stage of RF ampli-

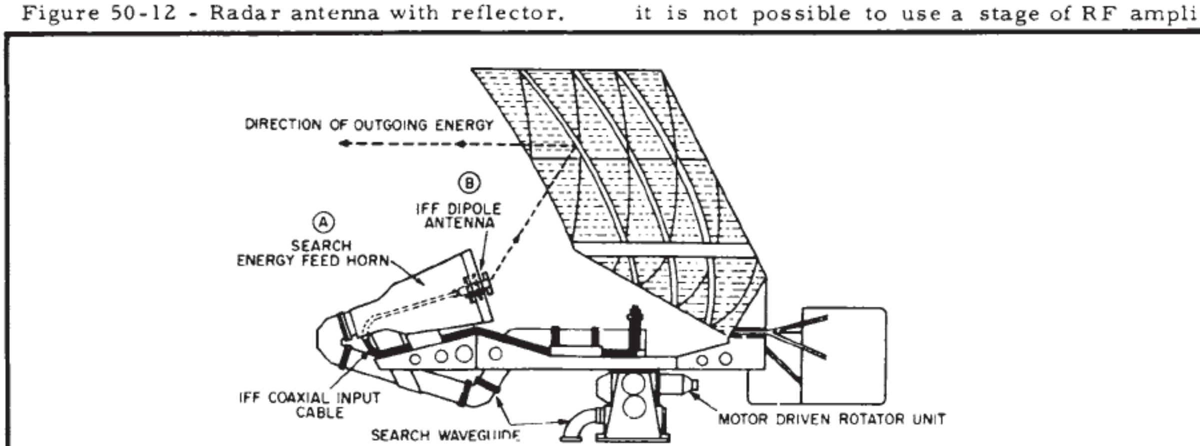


Figure 50-13 - This section of reflector dish is fed from two kinds of energy-transfer devices, (A) Energy feedhorn and (B) dipole antenna.

A14. The direction of electron travel is perpendicular to the magnetic field.

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cation ahead of the mixer, and therefore, the RF signals are fed directly to the mixer. This converter is shown in Figure 50-15.

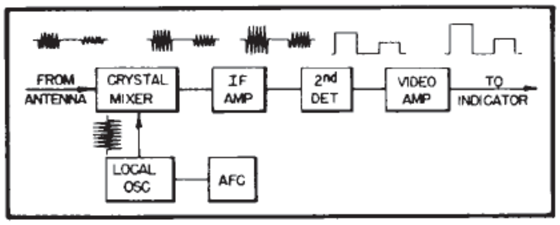


Figure 50-14 - Block diagram of a radar receiver.

In order to keep radar receivers in tune with their companion transmitters, a system of automatic frequency control is used in the receivers. Briefly, the system functions as follows: A small fraction of the RF energy from the transmitter line is applied to a special automatic-frequency-control (afc) mixer along with a small fraction of the RF energy from the receiver local oscillator. The IF energy resulting from the mixing of these two frequencies is amplified,

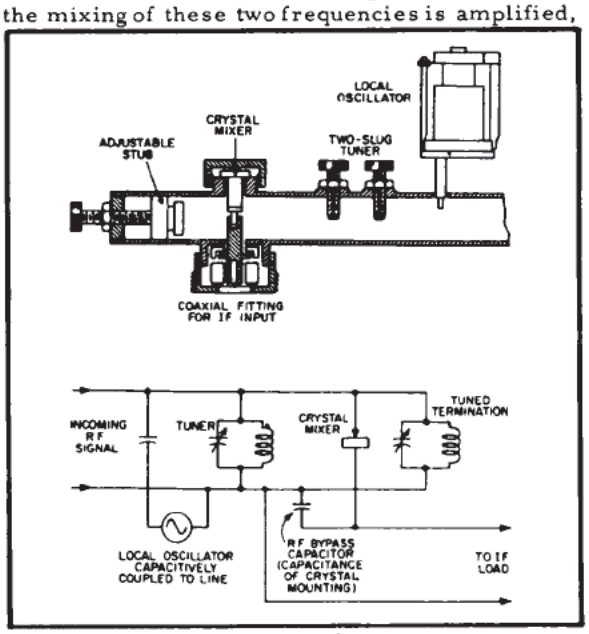


Figure 50-15 - Waveguide frequency converter which is part of the receiver equipment.

rectified, and applied via control circuits in such a way as to tune the local oscillator. If the IF is of the correct frequency, the resulting direct voltage maintains the local oscillator at the correct oscillator frequency. If the IF is too low in frequency the direct voltage applied to the local oscillator causes it to shift in frequency so that the IF will be increased. If the IF is too high, the oscillator frequency is shifted in the opposite direction.

The stability of operation is maintained in the microwave range of frequencies by careful design; and the overall sensitivity of the receiver is greatly increased by the use of many IF stages. Special types of tubes having low interelectrode capacitances also have been developed for use in local-oscillator and IF stages. Re-examine the block diagrams of a radar receiver which is shown in Figure 50-14. As in communications receivers, the IF signals in a radar receiver are fed to the second detector where the signal is rectified and the IF component is removed. The remaining modulation pattern, consisting of dc pulses, is fed to a video amplifier. In one type of presentation the output of the video amplifier is fed to the vertical deflection plates of an electrostatic-type cathode-ray tube. The amplitude of the vertical trace formed on the screen is proportional to the strength of the received signals. Simultaneously, a sawtooth voltage is applied to the horizontal deflection plates in synchronism with the transmitted pulse. The sawtooth voltage provides a horizontal displacement that is proportional to range.

Radar video amplifiers have wide band frequency response similar to that of television video amplifiers.

## 50-26. Radar Indicator

The radar indicator has the important function of converting the electrical output from the receiver into a visual display of range and bearing. Since the display is normally presented on a cathode ray tube, the indicator unit contains many of the circuits commonly found in oscilloscopes such as sweep oscillators, deflection amplifiers, and signal amplifiers (video amplifiers). When range information is to be obtained from the presentation, the indicator must also include the necessary range mark generator circuitry.

## 50-27. Radar Power Supply

In the functional diagram of the radar system (Figure 50-9) the power supply is represented as a single block. Functionally, this block is representative; however, it is unlikely that any one power supply could meet all the power requirements of a radar set. The distribution of the physical components of the system may be such as to make it impractical to lump the power-supply circuits into a single physical unit. Thus, different supplies are needed to meet the varying requirements of the system and must be designed accordingly. The power-supply function

is performed, therefore, by various types of supplies distributed among the circuit components of the radar equipment.

## RADAR SYSTEM CONSTANTS

Any radar system has associated with it certain specifications such as CARRIER FREQUENCY, PULSE-REPETITION FREQUENCY (the number of pulses sent out per second), PULSE WIDTH (in microseconds), and POWER RELATION (relationship of peak and average power). The choice of these arbitrary constants for a particular system is determined by its tactical use, the accuracy required, the range to be covered, the practical physical size, and the problem of generating and receiving the signal.

The travel time of a radar pulse is determined by the constant velocity of propagation of electromagnetic energy. The actual time required for a radar pulse to travel one nautical mile, strike a reflecting object, and return is 12.36 microseconds and the actual distance traveled is 2 x 2027 yards. In practice, the nautical mile is considered to be 2000 yards and the time required for the radar pulse to travel 2000 yards is 6.1 microseconds. The definitions of the radar mile and the nautical mile, and the time required for a pulse to travel a defined distance, varies from text to text. For calibration purposes the radar mile will be considered 2000 yards and the round trip distance propagation time 12.2 microseconds.

### 50-28. Carrier Frequency

The carrier frequency is the frequency at which the RF energy is generated. The principal factors influencing the selection of the carrier frequency are the desired directivity and the generation and reception of the necessary microwave RF energy.

For the determination of direction and for the concentration of the transmitted energy so that a greater portion of it is useful, the antenna should be highly directive. The higher the carrier frequency, the shorter the wavelength will be. Hence the antenna array is smaller for a given sharpness of pattern, because the individual radiating element is normally a half-wave long. For an antenna array of a given physical size the pattern is sharper for a higher frequency.

The problem of generating and amplifying reasonable amounts of RF energy at extremely high frequencies is complicated by the physical construction of the tubes to be used. The common triode becomes impractical and must be replaced by tubes of special design.

In general, the modifications for extremely high-frequency operation are designed to reduce interelectrode capacitances, transit time, and stray inductance and capacitance in the tube leads.

At the receiver end, it is very difficult to amplify microwave signals; as a result, RF amplifiers are not employed. Instead, the frequency of the incoming signal is mixed with that of a local oscillator in a crystal mixer to produce a difference frequency called the INTERMEDIATE FREQUENCY (IF). The intermediate frequency is low enough to be amplified in suitable IF amplifier stages employing electron tubes.

### 50-29. Pulse Repetition Frequency

Sufficient time must be allowed between each transmitted pulse for an echo to return from any target located within the maximum workable range of the system. Otherwise, the reception of the echoes from the more distant targets would be obscured by succeeding transmitted pulses.

The range of a radar set depends upon the pulse repetition rate provided the power is sufficient. For example, when the peak power is sufficient, and the repetition rate is 250 PPS, the period will be $\frac{10^6}{250} = 4000\ \mu s$. At 12.2 $\mu s$ per mile, the range will be $\frac{4000}{12.2} = 328$ miles. This necessary time interval fixes the highest pulse-repetition frequency that can be used to avoid interference with the returning echo by the next output pulse.

When the antenna system is rotated at a constant speed, the beam of energy strikes a target for a relatively short time. During this time, a sufficient number of pulses of energy must be transmitted in order to return a signal that will produce the necessary indication on the oscilloscope screen. For example, an antenna rotated at 6 rpm having a pulse repetition frequency of 800 cps will produce approximately 22 pulses for each degree of antenna rotation. The persistence of the screen and the rotational speed of the antenna therefore determine the lowest pulse repetition frequency that can be used.

### 50-30. Pulse Width

The minimum range at which a target can ideally be detected is determined largely by the width of the transmitted pulse. If a target is so close to the transmitter that the echo is returned to the receiver before the transmitter is turned off, the reception of the echo obviously will be masked by the transmitted pulse. For example, a pulse width of 1$\mu$s will have a minimum range of 164 yards, meaning that a target within this range will not show, or will be "blocked out" on the screen. In this respect, equipments for

A15. The maximum range desired.

A16. The pulse width.

---

"close in" ranging or navigation work use pulses of the order of 0.1 μ s. For long-range equipment the pulse width is normally from 1μ s to 5μ s.

Q15. What determines the PRF of a radar?

Q16. What is one factor that determines the minimum range of a radar set?

### 50-31. Power Relation

A radar transmitter generates RF energy in the form of extremely short pulses and is turned off between pulses for comparative long intervals. The useful power of the transmitter is that contained in the radiated pulses and is termed the PEAK POWER of the system. Power is normally measured as an average value over a relatively long period of time. Because the radar transmitter is resting for a time that is long with respect to the operating time, the average power delivered during one cycle of operation is relatively low compared with the peak power available during the pulse time.

A definite relationship exists between the average power dissipated over an extended period of time and the peak power developed during the pulse time. The overall time of one cycle of operation is the reciprocal of the pulse repetition frequency (PRF). Other factors remaining constant, the greater the pulse width the higher will be the average power; and the longer the pulse-repetition time, the lower will be the average power. Thus,

$$\frac{\text{Average power}}{\text{peak power}} = \frac{\text{pulse width}}{\text{pulse-repetition time}}$$

These general relationships are shown in Figure 50-16.

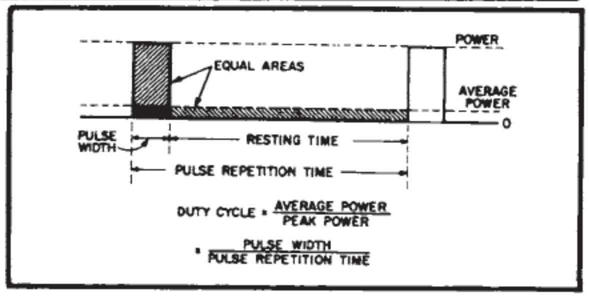


Figure 50-16 - Relationship of peak and average power

The operating cycle of the radar transmitter can be described in terms of the fraction of the total time that RF energy is radiated. This time relationship is called the DUTY CYCLE and may be represented as

$$\text{duty cycle} = \frac{\text{pulse width}}{\text{pulse-repetition time}}$$

For example, the duty cycle of a radar having a pulse width of 2 microseconds and a pulse-repetition frequency of 500 cycles per second (pulse repetition time = $\frac{10^6}{500}$, or 2,000 microsecond) is

$$\text{duty cycle} = \frac{2}{2,000} = 0.001$$

Likewise, the ratio between the average power and peak power may be expressed in terms of the duty cycle. Thus,

$$\text{duty cycle} = \frac{2 \times 10^{-6}}{2000 \times 10^{-6}}$$

In the foregoing example it may be assumed that the peak power is 200 kilowatts. Therefore, for a period of 2 microseconds a peak power of 200 kilowatts is supplied to the antenna, while for the remaining 1998 microseconds the transmitter output is zero. Because

average power = peak power x duty cycle.

average power = 200 x 0.001 = 0.2 kilowatts

High peak power is desirable in order to produce a strong echo over the maximum range of the equipment. Low average power enables the transmitter tubes and circuit components to be made smaller and more compact. Thus, it is advantageous to have a low duty cycle. The peak power that can be developed is dependent upon the interrelation between peak and average power, pulse width and pulse-repetition time, or duty cycle.

## EXERCISE 50

1. Why is it necessary to utilize AFC circuits in radar receivers?
2. Why aren't RF amplifiers found in radar receivers?
3. A radar transmitter has a peak power of 150 kilowatts and a duty cycle of 0.002. What is the average power?
4. A radar transmitter has a PRF of 800 cps, a resting time of 1249 microseconds, and an average power of 400 watts. What is the peak power?
5. The slant range of an aircraft is 5 miles. The angle of elevation of the radar antenna is 30°? What is the altitude of the aircraft?
6. What type of pulse is applied to a magnetron? To which element is the pulse normally applied?
7. Why is an electronic switch necessary in the antenna system of a radar which uses the same antenna for both sending and receiving?
8. What is the advantage of using a radar which has a low duty cycle?
9. What type of radar system performs a measurement of the difference in frequency between the transmitted and reflected energy?
10. Why does a surface search radar utilize narrow pulse widths?

# CHAPTER 51

# RADAR TIMER AND MODULATOR

This chapter will deal with the MASTER TIMER and MODULATOR of a radar set. The master timer, which is considered to be the heart of a radar set, produces accurately timed pulses, which are applied to both the transmitter and indicators. The modulator is used to produce high voltage pulses of the proper amplitude polarity, and width to trigger the radar transmitter.

## 51-1. Master Timer

The master timer unit supplies triggers to the radar indicators and transmitter. This unit must be very stable since it determines both the PULSE REPETITION FREQUENCY (PRF) and PULSE REPETITION TIME (PRT) of the radar set. The block diagram of a typical timer unit is illustrated in Figure 51-1.

The oscillator section generates a steady output at a fixed frequency. The frequency of operation is generally less than 1,000 cps, and will be determined by the purpose and type of radar set used. A phase shift or Wein-bridge oscillator is usually used to meet the requirements of stability and low frequency operation. These circuits are described in Chapter 46. The output of the oscillator section is a sine wave of constant amplitude and frequency.

The output of the oscillator is applied to an overdriven amplifier which produces a square wave output. The overdriven amplifier is discussed in section 48-18. This square wave should have steep leading and trailing edges to accurately preserve the time relationships produced by the oscillator. The square wave output of the overdriven amplifier is applied to an RC differentiating circuit. The differentiator produces two narrow pulses; a positive pulse corresponding to the $0^{\circ}$ point of the square wave, and a negative pulse corresponding to the $180^{\circ}$ point of the square wave.

The output is applied to a diode clipper which usually removes the negative pulse but in some cases positive clipping may be employed. The clipper output is applied to a cathode follower stage. The cathode follower stage is generally used to isolate the timer unit from other circuits in the set, and to provide impedance matching between the timer unit and the interconnecting coaxial cables. The output of the timer unit consists of very narrow pulses, usually 30 to 60 volts in amplitude, which correspond to the $0^{\circ}$ point of each cycle produced by the timer oscillator.

## 51-2. Radar Modulators

The modulator section, which is usually triggered by the master timer, produces a high negative pulse, which is used to fire the master oscillator power amplifier (Magnetron), in the radar's transmitter section. The amplitude of this pulse will be determined by the type of master oscillator power amplifier (Magnetron) used. Pulse width is critical since it determines the length of time the transmitter will be producing an output. The pulse width may be a fixed value of from 0.1 to 10 micro seconds, and will be dependent on the frequency and application of a particular radar set.

The block diagrams of two types of modulator systems are illustrated in Figure 51-2. The LINE PULSING MODULATOR (Figure 51-2A consists of the following components:

1. The source - Provides high voltage to the storage element in the modulator.

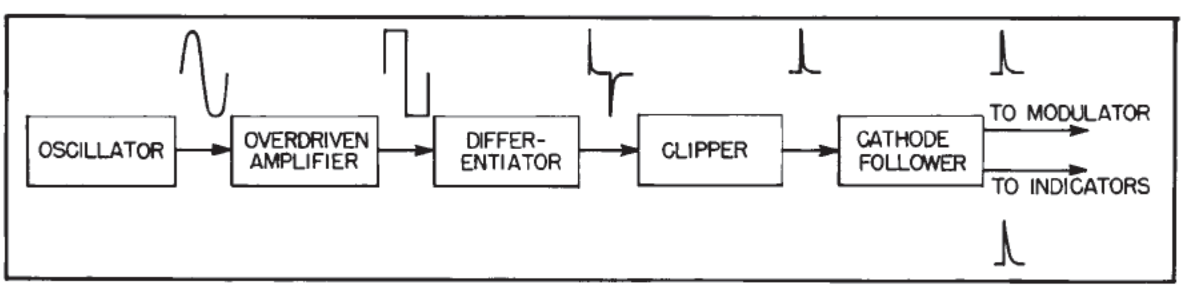


Figure 51-1 - Master timer unit.

2. $Z_{ch}$ - The total impedance in the storage network charge path.
3. The Switch - (May be a spark gap or thyratron) allows the storage network to charge when open; allows the storage network to discharge when closed.
4. PULSE FORMING NETWORK - Stores the high voltage provided by the source. Also determines the shape and width of the HV pulse applied to the load.
5. Load - Primary winding of the pulse transformer.

The DRIVER HARD TUBE MODULATOR (Figure 51-2B) differs from the line pulsing type in many respects. Although $Z_{ch}$, the source, and the load are similar to the line pulsing type; the storage network in the hard tube system does not shape the output pulse. The hard tube modulator (Figure 51-2B) consists of the following components:

1. $Z_{ch}$, Source, Load - Same as in the line pulsing type.
2. STORAGE NETWORK - Usually a large capacitor. Does NOT determine pulse width.
3. HARD TUBE SWITCH - Controls the charge and discharge of the storage unit. When the switch tube is cut off, the storage unit charges; when the switch tube conducts, the storage unit discharges.
4. DRIVER - Input is a trigger from the master timer; output is a HV pulse used to drive the switch tube into conduction. Since the width of this pulse determines how long the switch tube will conduct, the driver pulse width will determine the width of the HV pulse applied to the load.

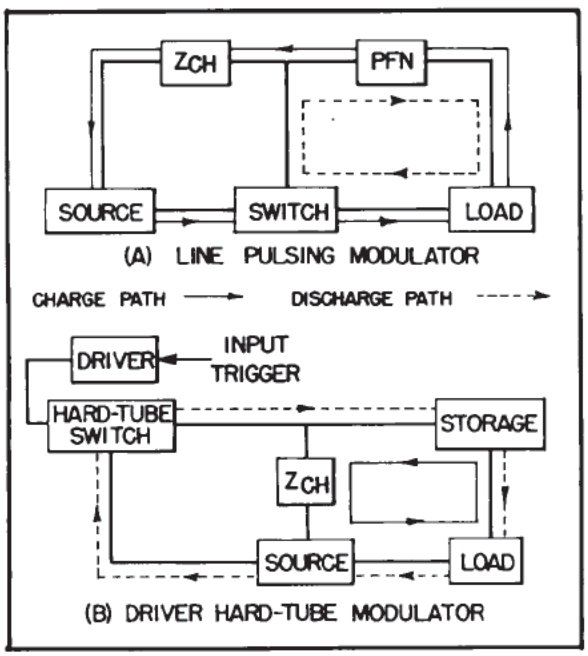


Figure 51-2 - Two types of modulators.

Q1. What section determines the pulse shape in a line pulsing type modulator?

## 51-3. Artificial Transmission Lines

Some of the characteristics of a transmission line, such as time delay, pulse shaping, and energy storage, can be used to advantage in a radar set. Transmission lines are discussed in detail in Chapter 27. The physical length of a real transmission line required for a particular application may be too great for practical use. In this case an ARTIFICIAL TRANSMISSION LINE (designated hereafter as line) may be constructed by first determining the values of L and C in the real line, and lumping these component values into an artificial line. The line will then have the same electrical characteristics as the real line. The construction of a line is illustrated in Figure 51-3. Note that lumped values of L and C are used to replace the distributed L and C values which occur in a real line.

The artificial line will produce the same delay characteristics as a real line. A voltage applied to the line terminals at point A will appear at the terminals at point B delayed by a time interval which is determined by the values of L and C. The time delay may be calculated by using the formula:

$$Td = N\sqrt{LC} \qquad (27\text{-}4)$$

where: T is the delay time in seconds, N is the number of line sections; L the inductance in henrys per section, and C the capacity in farads per section.

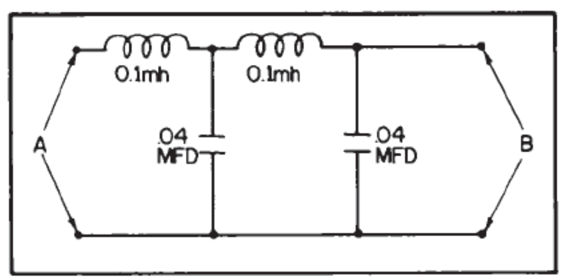


Figure 51-3 - Artificial transmission line construction.

The line illustrated in Figure 51-3 will have a delay time of:

$$Td = N\sqrt{LC} = 2\sqrt{1 \times 10^{-4} \times 4 \times 10^{-8}}$$
$$= 2\sqrt{4 \times 10^{-12}}$$
$$= 2 \times 2 \times 10^{-6}$$
$$= 4 \times 10^{-6} \text{ seconds or 4 micro seconds}$$

An EMF applied to the input terminals of this line will NOT appear at the output terminals until 4 micro seconds have elapsed.

Since the line presents the same series-parallel RLC circuit characteristics as a real line; the method of determining the characteristic impedance developed in Chapter 27 may be used. The formula for determining characteristic impedance is:

$$Z_o = \sqrt{\frac{L}{C}} \qquad (27\text{-}1)$$

Where $Z_o$ is the characteristic impedance in ohms; L is the inductance per section in henrys; and C is the capacity per section in farads. The $Z_o$ of the circuit illustrated in Figure 51-3 is:

$$Z_o = \sqrt{\frac{L}{C}} = \sqrt{\frac{1 \times 10^{-4}}{4 \times 10^{-8}}} \qquad (27\text{-}1)$$
$$= \sqrt{0.25 \times 10^{4}}$$
$$= \sqrt{25 \times 10^{2}}$$
$$= 5 \times 10 \text{ or 50 ohms}$$

Note that the addition of more line sections will not change the $Z_o$ since the L/C ratio remains constant.

The line is similar to the two-wire transmission line, and does have a high frequency limitation. Since the line is composed of inductance and capacitance; filtering action will take place. At a very high frequency, the square wave pulse will become distorted. For a complete analysis of this type of distortion refer to Chapter 45.

Q2. What is the total delay time of an open ended, three section line, which contains 2 millihenry of inductance and 0.008 micro farads of capacitance per section?

## 51-4. Charging Open Ended Lines

The dc charging circuit of an open ended artificial transmission line is illustrated in Figure 51-4A. $R_{ch}$ is the internal resistance of the source. The $Z_o$ of the line can be found by the formula:

$$Z_o = \sqrt{\frac{L}{C}} = \sqrt{\frac{1 \times 10^{-4}}{1 \times 10^{-8}}} = \sqrt{1 \times 10^{4}} = 100 \text{ Ohm}$$

The time delay (Td) for one way travel is:

$$Td = N\sqrt{LC} = 2\sqrt{1 \times 10^{-4} \times 1 \times 10^{-8}}$$
$$= 2 \times 10^{-6} \text{ or 2 micro seconds}$$

The dc charging sequence starts when $S_1$ is closed. Since $Z_o = R_{ch}$, one half of $E_{bb}$ will be applied to the line at point (1). The first section of the line, composed of $L_1$ and $C_1$, charges to $E_{bb}/2$ (50v) at $T_d/2$ or 1 micro second (Figure 51-4B). At this time 50v is measured at points (1) and (2), but the voltage at point (3) is zero. Two micro seconds after $E_{bb}$ is applied to the circuit, the second section of the line, $L_2$ and $C_2$, charges to $E_{bb}/2$ or 50v. At this instant 50v will be measured at points (1), (2) and (3). The voltage wave which travels from the input terminals to the output terminals of the line is called THE INCIDENT WAVE.

While the line is charging to $E_{bb}/2$, charge current flows through $L_1$ and $L_2$. The instant $C_2$ charges to 50v, the magnetic field around $L_2$ collapses, increasing the charge on $C_2$ to $E_{bb}$ or 100v. At this instant the voltage measured at point (3) is 100v; this is the start of the REFLECTED WAVE. The open end of the line now acts as a 100v source, and the reflected wave travels back to the input terminals, increasing

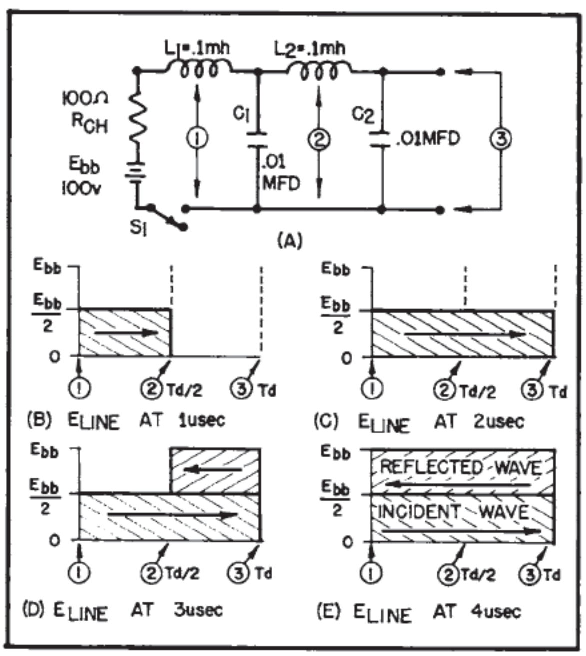


Figure 51-4 - DC charging circuit for an open-ended ATL.

A1. Pulse shape is determined by the pulse forming network.

A2. Total time delay $T = 2N\sqrt{LC}$ Then,

$$T = 2 \times 3\sqrt{2 \times 10^{-3} \times 8 \times 10^{-9}}$$
$$= 6\sqrt{16 \times 10^{-12}} = 6 \times 4 \times 10^{-6}$$
$$= 24 \times 10^{-6} \text{ seconds or 24 micro seconds}$$

---

the charge on the line from 50v to 100v. After 3 micro seconds 50v is measured at point (1) while 100v is measured at points (2) and (3) (Figure 51-4D). After 4 micro seconds 100v is measured at points (1), (2) and (3), and the line is fully charged to $E_{bb}$. Note that the total time required to charge the line to $E_{bb}$ is twice the one way travel time or 2Td. A charged open end line will have maximum voltage (equal to $E_{bb}$) and zero current.

It is necessary that $Z_o$ and $R_{ch}$ be equal if the line is to be fully charged after one incident and reflected wave. If $R_{ch}$ is greater than $Z_o$, the voltage applied to the line is less than $E_{bb}/2$, and several incident and reflected waves will be required to charge the line to $E_{bb}$. If $R_{ch}$ is less than $Z_o$, the voltage applied to the line is greater than $E_{bb}/2$, and the line will oscillate above and below the value of $E_{bb}$ several times before charging to $E_{bb}$.

A shorted end line will charge in the same manner as the line in Figure 51-4. In the shorted end line, the incident and reflected waves will be current waves rather than voltage waves. When the shorted end line is fully charged, current will be maximum and voltage will be zero.

## 51-5. DC Resonance Charging

Another method of charging an artificial transmission line (line) is illustrated in Figure 51-5A. In this circuit, the artificial line will be used for charge storing and designated $C_{st}$. Charge current flows through a series inductor $L_{ch}$. The inductance of $L_{ch}$ is much greater than the inductance present in the $C_{st}$. When $S_1$ is closed, $C_{st}$ starts to charge to $E_{bb}$. The charge current of $C_{st}$ flowing through $L_{ch}$ produces a magnetic field. When $C_{st}$ is charged to $E_{bb}$, the field around $L_{ch}$ collapses, increasing the charge on $C_{st}$ to approximately 1.9 $E_{bb}$ (Figure 51-5B). At this point, the field surrounding $L_{ch}$ has completely collapsed, and $C_{st}$ will discharge through the source. Since $C_{st}$ and $L_{ch}$ forms a series resonant circuit, the voltage appearing across $C_{st}$ will take the form of a damped sine wave with $E_{bb}$ as the reference level. The frequency of oscillation will be dependent on the values of $L_{ch}$ and $C_{st}$; the damping time will be

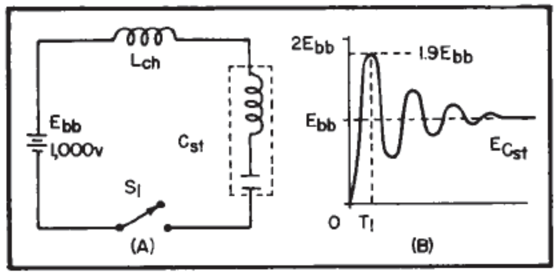


Figure 51-5 - DC resonance charging circuit.

dependent on the circuit $I^2R$ losses.

The advantage of this charging method lies in the magnitude of $C_{st}$ charge occurring at $T_1$ in Figure 51-5B. If, at this instant, $C_{st}$ is disconnected from the charge path and connected to a load, it can supply a voltage which is 95% greater than the source $E_{bb}$. This system would require very critical timing. If $S_1$ were automatically opened at $T_1$, $C_{st}$ would retain its charge of 1.9 $E_{bb}$.

In Figure 51-6A the switch is replaced by diode $V_1$. $C_{st}$ charges through $V_1$ and $L_{ch}$. When $C_{st}$ is charged to $E_{bb}$, the field around $L_{ch}$

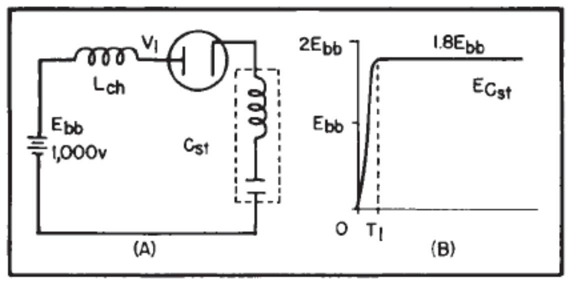


Figure 51-6 - DC resonance charging with diode.

collapses, charging $C_{st}$ to about 1.8 $E_{bb}$. The charge on $C_{st}$ is less than the charge achieved in Figure 51-5B due to the voltage drop across $V_1$. $C_{st}$ cannot discharge due to diode $V_1$, and will now retain a charge of 1.8 $E_{bb}$ until a discharge path is provided.

Q3. What are the advantages of charging an artificial transmission line through a charging choke and diode?

## 51-6. Discharging Open Ended Lines

Figure 51-7 illustrates the waveshape produced when an artificial transmission line is discharged into a load which has an impedance ($Z_R$) that matches the characteristic impedance ($Z_o$) of the line. Assume that the line ($Z_o$) had

been charged to some value of voltage ($E_{bb}$) and $S_1$ is open. At this time line voltage equals $E_{bb}$ and load voltage equals zero. $S_1$ is closed at time $T_o$ (Figure 51-7B). At this instant, line voltage ($E_{Z_o}$) drops to $E_{bb}/2$ and load voltage ($E_{Z_R}$) increases to $E_{bb}/2$. The changes are equal because $Z_o = Z_R$. The discharge sequence

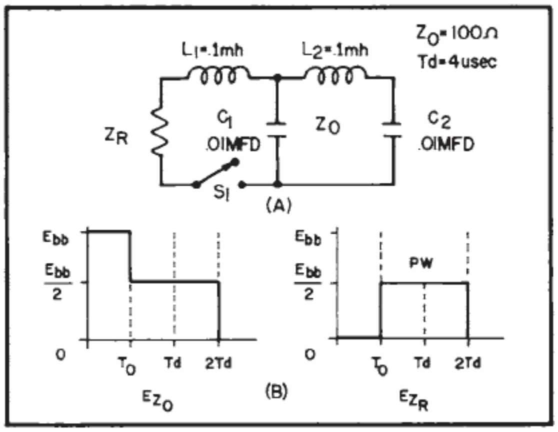


Figure 51-7 - Discharging a line when $Z_o = Z_R$.

of the line is similar to the charge sequence. An incident wave occurs from time $T_o$ to $T_d$ discharging the line by half. The reflected wave occurs from time $T_d$ to $2T_d$ completely discharging the line. When $Z_R = Z_o$, the output voltage ($E_{Z_R}$) will be a pulse with an amplitude equal to $E_{bb}/2$ and a pulse width (PW) equal to $2T_d$.

When load impedance is greater than line impedance, several incident and reflected waves will occur before the line is completely discharged. The pulse which will appear across the load impedance is illustrated in Figure 51-8. Note that the pulse width is greater than the

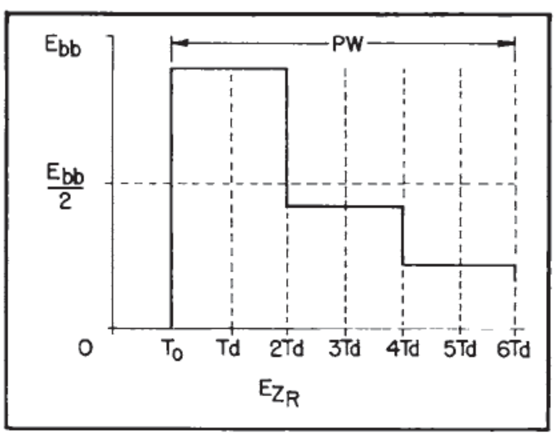


Figure 51-8 - Pulse voltage when $Z_R$ is greater than $Z_o$.

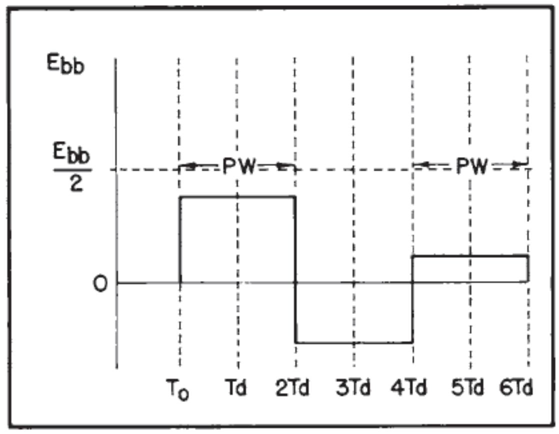


Figure 51-9 - Pulse voltage when $Z_R$ is less than $Z_o$.

normal time of 2Td.

When load impedance is less than line impedance, the load voltage is less than the line voltage and after a discharge time of 2T, the line recharges in the opposite polarity. The resultant output pulse is illustrated in Figure 51-9. Note that once again the total discharge time is greater than 2Td.

Artificial transmission lines have many applications in radar transmitters and receivers. In receivers they are used to provide controllable time delays for triggers and video signals. In transmitters they are used as pulse forming networks (line) to store and deliver high voltage energy to the form of a pulse to the transmitter's master oscillator power amplifier (Magnetron).

## 51-7. Switching Devices

After the pulse forming network has been charged, some method of switching must be provided to allow the pulse forming network to discharge through the load. The switch must be able to handle a high pulse repetition frequency (PRF) and pass a current flow which may exceed 100 amperes. The switch must also provide accurate control of the on and off time interval.

One method of switching is illustrated in Figure 51-10A. An arc is produced each time one of the rotating electrodes passes in close proximity to the fixed electrode. The number of rotating electrodes and speed of rotation will determine the number of arcs (switch closures) per second. For instance, if the speed of rotation was 6,000 revolutions per minute (100 revolutions per second) and the number of electrodes is four, the switch would effectively open and close 400 times per second.

A3. The advantages of charging an artificial transmission through a charging choke and diode are; the line will charge to a voltage that is 90% greater than that applied by the source; and the line switching time for discharge is not critical.

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A modulator circuit utilizing a rotary spark gap switch is illustrated in Figure 51-10B. The pulse forming network charges to source voltage when the fixed electrode is between rotating electrodes 1 and 2. The pulse forming network discharges through the load when there is an arc between the fixed electrode and rotating electrode number 2. The cycle keeps repeating between each set of rotating electrodes. The rotary spark gap is seldom used because of its short life expectancy and unstable firing time. Firing time may vary because of electrode corrosion, changes in air temperature, pressure, and humidity.

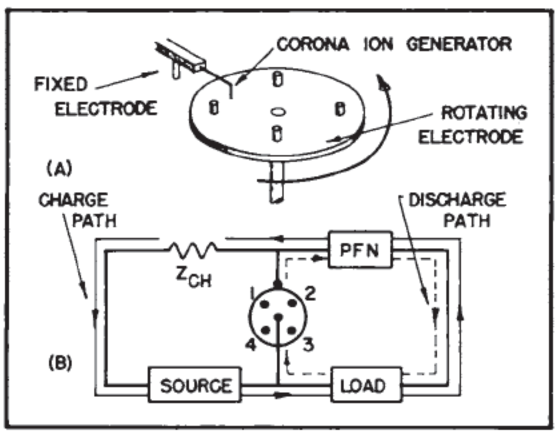


Figure 51-10 - (A) Rotary spark gap (B) Circuit using rotary spark gap.

A better method of switching is illustrated in Figure 51-11A. In this case, a hydrogen thyratron is used as the switching device. During $T_1$ (Figure 51-11B) the thyratron is not conducting, and the pulse forming network (line) charges through $L_{ch}$ and the charging diode. The pulse forming network charges to approximately 1.8 times the source voltage. At the start of $T_2$, a trigger pulse, which corresponds in time with the master timer trigger, is applied to the grid of the thyratron. The thyratron is driven into conduction, allowing the line to discharge into the load. Since the line impedance ($Z_o$) of the line is closely matched to the impedance of the load $Z_R$, one half of the line voltage is dropped across the load.

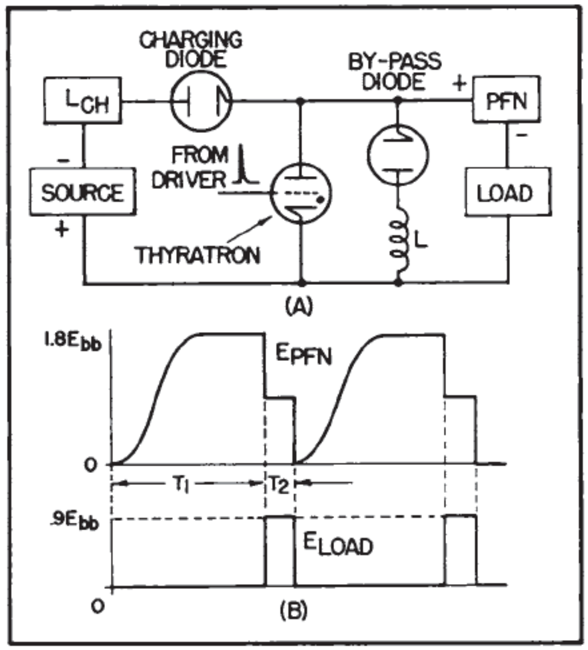


Figure 51-11 - (A) Modulator circuit using a thyratron switch.
(B) Line and load voltages for two cycles of operation.

At the end of $T_2$, the line voltage decreases rapidly and the thyratron de-ionizes. The width of the pulse appearing across the load is determined by the two way delay time of the line.

If there is a slight impedance mismatch ($Z_R$ is less than $Z_o$) between the pulse forming network and the load, the line will tend to recharge slightly in the opposite direction (Figure 51-9). This opposite charge would increase the current through $L_{ch}$, causing the pulse forming network voltage to increase during the normal charge cycle. Pulse forming network voltage will keep increasing during each succeeding cycle until breakdown of either the load or thyratron occurred.

The by-pass diode and inductor L are connected in shunt with the thyratron to prevent inverse charging of the line. If an inverse charge is present on the line, the by-pass diode conducts providing a discharge path through the inductor. The discharge current produces a magnetic field around $L_1$. When the discharge current stops, the field around $L_1$ collapses causing a slight charge of the proper polarity in the pulse forming network.

Q4. When a thyratron is used as the switch tube in a modulator, why must a pulse forming net-

work be used, rather than the trigger pulse, to control output pulse width?

Q5. What is the purpose of the by-pass diode connected in shunt with the thyratron switch?

## 51-8. Pulse Transformer

The load connected to the line or storage circuit, will be applied to the pulse transformer of a magnetron (master oscillator power amplifier). A pulse transformer is used to step up the high voltage pulse from the pulse forming network and provide impedance matching between the magnetron and the line. Pulse transformer design is critical because of the high frequency components present in the output pulse. The core is composed of thin laminations of ferro-magnetic material, usually silicon steel. Close coupling between primary and secondary reduces leakage inductance to preserve the steep leading edge of the input pulse. Low interwinding capacitance is desired to prevent high frequency oscillations. Close coupling is attained between primary and secondary, by winding the primary directly on the secondary and by using the same leg of the core for both windings. The secondary is usually BIFILAR (meaning two) windings.

The bifilar secondary is illustrated in Figure 51-12. The secondary is made up of two insulated conductors, wound side by side so that exactly the same voltage is induced in each. The bifilar winding acts as two secondaries,

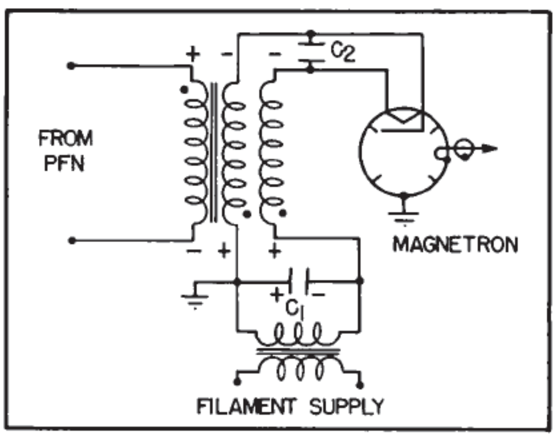


Figure 51-12 - Bifilar secondary.

which have equal and in-phase voltages induced in them. The bifilar winding permits the use of a filament secondary without high voltage insulation.

By-pass capacitors, $C_1$ and $C_2$, are often used so that pulse current will flow directly to the bifilar winding, without affecting the filament circuit.

Q6. What is the advantage of using a bifilar secondary winding on a high voltage pulse transformer that is connected to a magnetron?

## 51-9. Protective Devices

Occasionally an over voltage condition may exist in the pulse forming network. The line will charge to a higher voltage than normal, and an excessively high negative pulse will be applied to the magnetron. Frequent overvoltages can cause arcing and damage to the magnetron. An overvoltage spark gap is connected across the pulse transformer secondary to prevent excessively high pulse voltages from being applied to the magnetron (Figure 51-13A). The gap width is manually adjusted for the desired value of voltage which will produce arc-over.

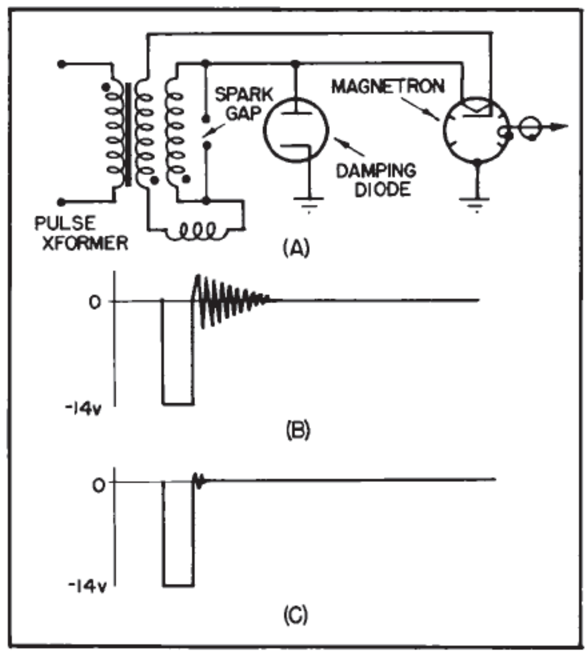


Figure 51-13 - Protective devices used with the magnetron.

Stray capacitance and leakage inductance in the pulse transformer secondary circuit, produces a series of oscillations after the main pulse has been applied (Figure 51-13B). The negative portions of this waveshape will produce a spurious output from the magnetron which can obliterate any short range targets. These oscillations are not produced during the main pulse, due to the low impedance shunting provided by the conducting magnetron. A DAMPING

A4. A line must be used to control output pulse width when a thyratron switch is used, because after ionization, the thyratron grid loses control, and a decrease in grid potential cannot cause the de-ionization necessary to end the output pulse.

A5. The by-pass diode prevents the line from charging to a polarity which is the inverse of normal.

A6. The bifilar secondary simplifies the insulation problem of the magnetron heater circuit.

---

DIODE is connected in parallel with the magnetron to eliminate the effects of these oscillations. When the negative main pulse is applied to the magnetron cathode, the damping diode does not conduct. The damping diode will conduct during the positive portion of the oscillations (Figure 51-13C); shunting the magnetron with a low impedance and causing the oscillations to dampen very rapidly.

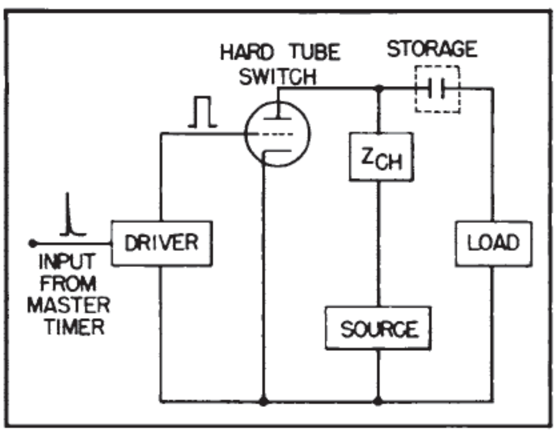


Figure 51-14 - Driver hard tube modulator.

## 51-10. Bootstrap Driver

A DRIVER HARD TUBE MODULATOR is illustrated in Figure 51-14. In this modulator, the storage unit is a capacitor. It will NOT determine the width or shape of the pulse. The conduction level, and conduction time of the hard tube switch will determine the width and shape of the pulse applied to the load.

The driver unit applies a constant amplitude pulse (produced by the timing trigger) to the hard tube switch, causing it to conduct. Since the driver pulse determines the conduction time of the hard tube switch, the width of the pulse applied to the load will be equal to the width of the driver pulse. Although other types of drivers are available, only the bootstrap driver will be discussed.

The simplified block diagram of a bootstrap driver is illustrated in Figure 51-15. Positive triggers from the master timer are applied to the grid of the thyratron through the ISOLATION DIODE. The isolation diode eliminates any adverse transient effect, which may influence the operation of the trigger circuit. The input trigger fires the thyratron, which allows the pulse forming network to discharge. The discharging line applies a positive 300v pulse to the bootstrap amplifier. Pulse width is determined by the line. The bootstrap amplifier is a cathode loaded amplifier, which produces an amplified output pulse that is in phase with the input. The output pulse has an amplitude of minus 1000v to plus 300v, or a total amplitude of 1300v.

A simplified schematic diagram of the bootstrap driver is illustrated in Figure 51-16. Before a trigger pulse is applied to $J_1$, the pulse forming network charges through resistors, $R_3$ and $R_4$, to 600v with the indicated polarity. At this time, $V_3$ has minus 1000v applied to its cathode and minus 1100v applied to its grid. The resultant bias of minus 100v holds $V_3$ in cut-off. The cathode voltage of $V_3$ (-1000 v) is also applied to the control grid of the hard tube switch in the modulator, holding it in cut-off.

When a positive trigger is applied to $J_1$, diode $V_1$ conducts, applying the trigger to the grid of thyratron $V_2$. The positive trigger causes $V_2$ to ionize, providing a discharge path for the pulse forming network. The line, which

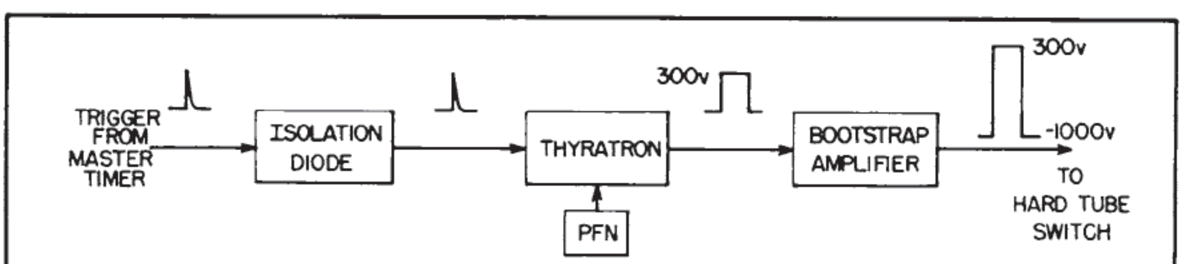


Figure 51-15 - Bootstrap driver-simplified block diagram.

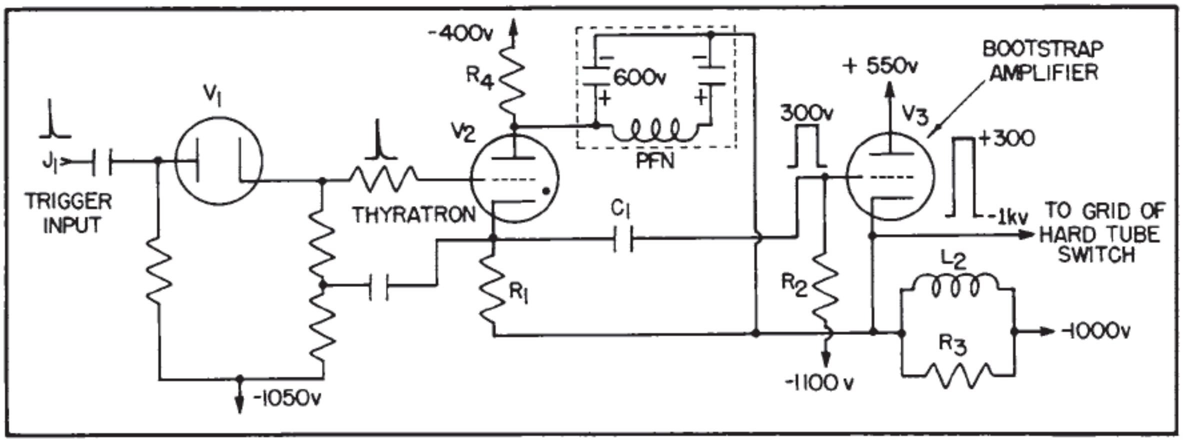


Figure 51-16 - Bootstrap driver - simplified schematic diagram.

was charged to 600v, discharges through $R_1$, producing a positive pulse of approximately 300v, to appear across $R_1$. This positive pulse is coupled to the grid of $V_3$ through $C_1$, driving $V_3$ into conduction. $V_3$ conducts heavily, and its cathode voltage rises from -1000v, to 300v. Output pulse amplitude is equal to the total change in cathode voltage, or 1300v. Pulse width is determined by the time delay of the pulse forming network.

The resultant waveshape appears at the grid of the hard tube switch as a positive 1300 volt pulse, superimposed on a negative 1000 volt bias level. This pulse drives the hard tube switch into conduction, which in turn, discharges the storage network in the modulator. A modulator pulse, equal in width to the driver pulse, is now applied to the magnetron.

## EXERCISE 51

1. Describe the function and operation of each block of the radar master timer unit.
2. What types of oscillators are normally used in the timer? Why are these types chosen?
3. What are the requirements of the square wave output from the overdriven amplifier?
4. What is the function of a radar modulator?
5. Why is the pulse width of the modulator pulse critical?
6. Describe the block diagram of the hard tube modulator?
7. Compare an artificial transmission line to an ordinary transmission line. Why is an artificial line used instead of an ordinary transmission line?
8. How much of a delay may be expected from an eight section line the inductance of one section of which is one millihenry and the capacitance is 0.005 microfarad? What is the characteristic impedance of the line?
9. Describe how an open-ended line is charged.
10. Describe the charging action of a shorted line.
11. What is resonance charging?
12. What is the purpose of a charging diode and a charging choke?
13. What is the disadvantage of the spark gap switching device? How is this disadvantage overcome by the use of a thyratron?
14. Describe the operation of a modulator circuit which uses a pulse forming network and a thyratron?
15. What is the purpose of a by-pass diode?
16. How does a pulse transformer differ from an ordinary power transformer?
17. What is the function of the bifilar winding?
18. Why is a damping diode required?
19. What is the bootstrap driver? What are its characteristics and advantages?
20. Why is the isolation diode used in the Bootstrap circuit?

CHAPTER 52

# WAVEGUIDES AND CAVITIES

As radar frequencies became increasingly higher, the open-wire line and coaxial line were no longer suitable as transmission lines because of their high energy loss. The WAVEGUIDE was developed to conduct high energy microwaves.

The purpose of a waveguide is to transfer energy from one point to another, in the same manner as an open wire or coaxial transmission line. However, waveguides differ from the open-wire or coaxial line because all of the transmitted energy is contained within the waveguide in the form of electric and magnetic fields rather than the conventional current and voltage movement. This controlled energy movement can be thought of as similar to directed wave propagation.

Waveguides are most commonly constructed in configurations similar to sectional pieces of hollow pipe or conduit. They can have either a circular or rectangular cross section area as shown in Figure 52-1.

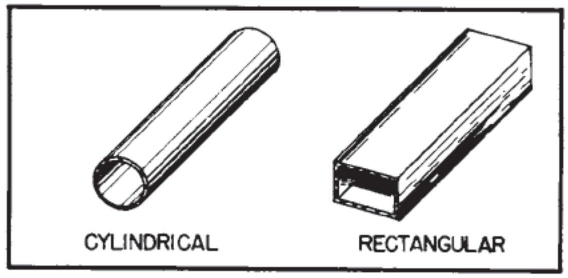


Figure 52-1 - Types of waveguides.

## 52-1. Development of Waveguides From Parallel Lines

Figure 52-2 shows a section of two-wire transmission line of the most simple construction supported on two insulators A and B. The insulators may be made of plastic, porcelain, or similar material.

From the view point of the line, insulator A is an impedance, $Z_1$, to ground; and insulator B is an impedance, $Z_2$, to ground. Of course these impedance values must be very high, otherwise the line would be shorted or bypassed to ground.

$Z_1$ and $Z_2$ are not necessarily pure resistances. The presence of the insulator causes a significant amount of capacitive reactance to be introduced into the system. A capacitor is formed between each conductor and ground. The dielectric for this capacitor is the insulator itself. However, the presence of capacitance as well as the resistance makes little difference in the usual installations as long as the total impedance is kept very high. It is desirable to keep the insulator losses as low as possible. Therefore, the insulator should have high leakage resistance and low dielectric loss.

Since the wires have impedance to ground and ground itself is a low impedance, the wires have an impedance between them. Therefore, the wires look upon the terminals of the insulators A and B as the terminals of a high impedance. $Z_3$ made up of the combined effects of the other impedances.

Because a quarter-wave line shorted at one end acts at the other end as a very high impedance, it can be used as an insulator. Since the transmission line in Figure 52-2 regards its insulators as two terminals between which a high impedance exists, the line can be very effectively supported on a quarter wave stub as shown in Figure 52-3. In fact, $Z_1$, $Z_2$ and $Z_3$ now are higher in value as compared to the more conventional insulators because a quarter-wave line has lower losses. This quarter-wave

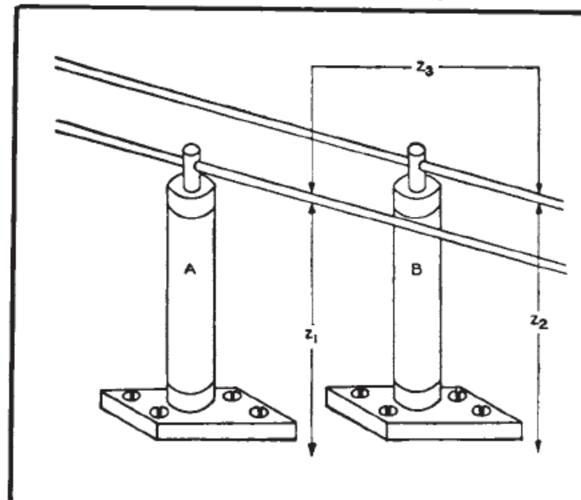


Figure 52-2 - Two wire transmission line using ordinary insulators.

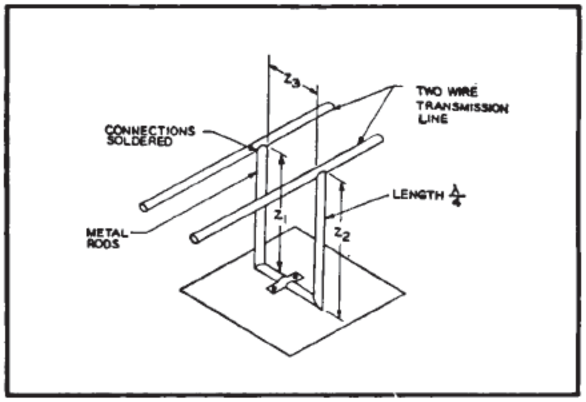


Figure 52-3 - Two wire transmission line using quarter-wave insulators.

line is sometimes called a METALLIC INSULATOR.

While the insulators in Figure 52-2 may be used for a wide range of frequencies, the quarter-wave line may be used for only ultra high frequencies and then for only a very narrow band of frequencies. If a widely different frequency is used, the stub length is no longer a quarter-wave length long, and therefore, it no longer can act as an insulator.

Of course, a metallic insulator such as the quarter-wave stub is an insulator only because its action as a parallel resonant circuit, which of necessity, is not practicable. Although limited, to ultra high frequencies and to very narrow bands, such insulators have the advantage of mechanical simplicity and unusually low power loss.

In Figure 52-4, a two-wire line is supported at both top and bottom by quarter-wave insulators. In effect, the line is placed at the center of a half-wave line shorted at both ends, which is called a half-wave frame. If the quarter-wave supports in Figure 52-5 are placed so close together that they touch at all points, a rectangular metallic tube is created. The original transmission lines now become a part of the side walls of the tube, and the top and bottom quarter-wave lines are the top half and the bottom half of the tube as shown in Figure 52-5B. This solid structure, called a waveguide, can be thought of as being composed of two bus bars and a multitude of quarter-wave insulators. Actually, for mechanical simplicity the tube is made of sheet metal, rather than of metal rods which are soldered together.

A waveguide constructed simply by increasing the number of half-wave frames until they touch carries not only one frequency, but also all higher frequencies. This can be explained when Figure 52-5 is compared with Figure 52-6.

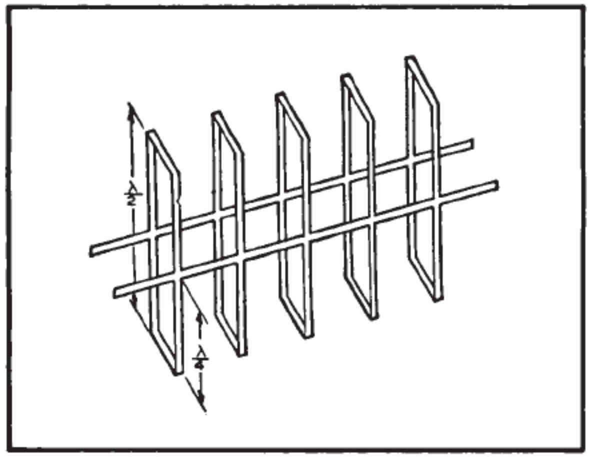


Figure 52-4 - Two-wire transmission line using double quarter-wave insulators.

A waveguide may be considered as having upper and lower quarter-wave sections of metallic insulation and a central section of bus bar. In Figure 52-5 the distance ab equals cd which equals one-quarter wavelength. The distance bc is the width of the bus bar. Assuming the dimensions of the waveguide are held constant, at some higher frequency, the width of the bus bar in Figure 52-6 in effect is increased to b'c', while the quarter-wave insulators decrease in length until a'b' equals c'd' which, in turn, equals a quarter wavelength at the new frequency. Theoretically, the waveguide could pass an infinite number of frequencies and the quarter wavelengths approach zero; and the bar occupies the entire side of the guide. In practice, this is limited by certain other factors.

One important fact should be noted. If the wavelength increases (frequency decreases) so much that the two quarter-wave insulators

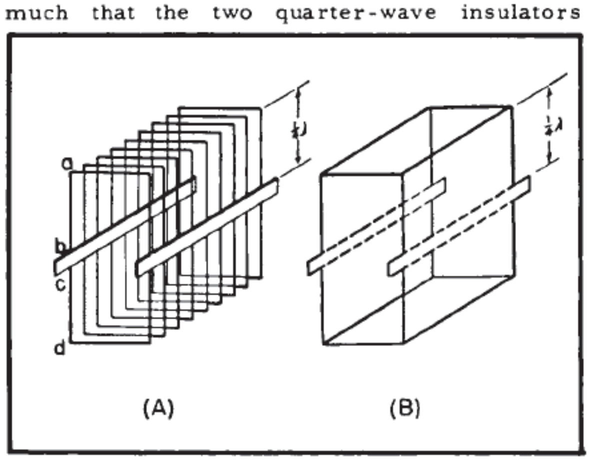


Figure 52-5 - Waveguide near minimum frequency.

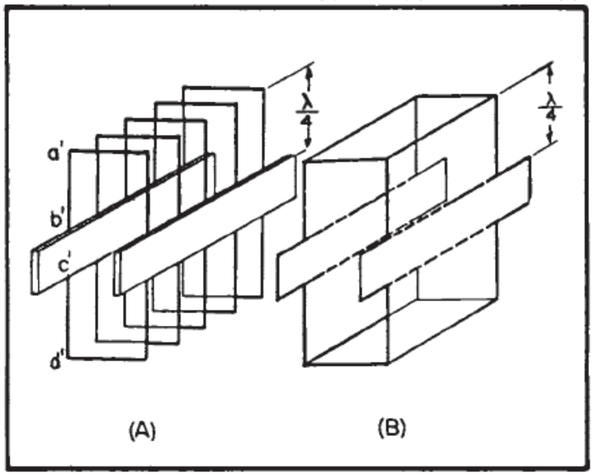


Figure 52-6 - Waveguide above minimum frequency.

CANNOT be created within the distance ab in Figure 52-7, the insulators automatically become less than a quarter wavelength. In this case, instead of being high resistive impedances; they become much lower inductive impedances; and the current is shorted out. Thus, a waveguide ceases the transmission of energy at the cutoff frequency, blocking any lower frequencies, but will transmit all frequencies above the cutoff frequency. The distance "ab" in figure 52-7 is between .51 to .84 wavelengths in most applications.

Q1. In a two-wire line suspended from the ground by insulators, why is there an impedance between the wires?

Q2. Why is a metallic insulator practicable?

Q3. What is the relationship between frequency and wavelength?

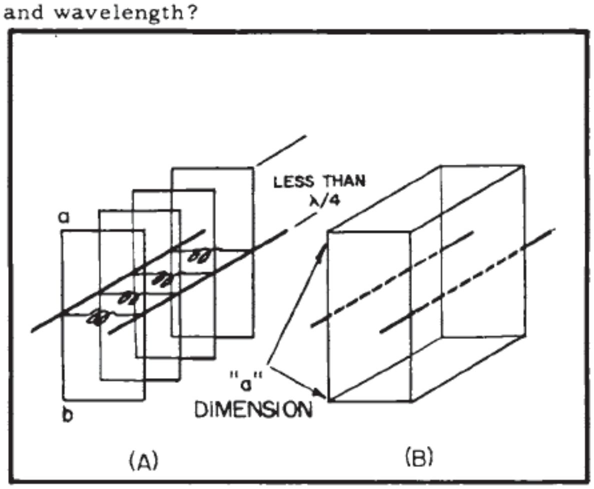


Figure 52-7 - Waveguide below minimum frequency.

## 52-2. Power Handling Capabilities of Waveguides

The maximum power that can be passed through a waveguide is directly proportional to the maximum voltage that can exist inside a waveguide. As in capacitors, the working voltage is dependent upon the distance between the two plates and the type of dielectric material as shown in Figure 52-8. This same condition holds true with waveguides. As shown in Figure 52-8, the narrow or "b" dimension will determine the voltage handling capacity.

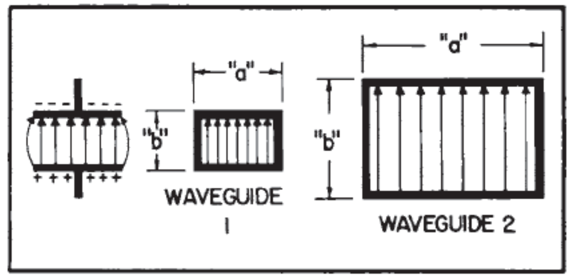


Figure 52-8 - Power dimensions of waveguides.

If the "b" dimension of waveguide 2 is twice that of waveguide 1, twice the value of voltage will be able to exist in waveguide 2. Since power is a function of the square of the voltage, the power handling capabilities of waveguide 2 would be four times greater than waveguide 1. The narrow or "b" dimension will determine the maximum power handling capacity of the waveguide.

## 52-3. Advantages of Waveguides

One of the prime reasons for the use of waveguides is the low attenuation of energy traveling through it. Loss in a transmission line can be divided into three categories: radiation loss, copper loss, and dielectric loss.

Radiation loss is the most serious in the open-wire line since there is no provision to contain any escaping energy. The coaxial cable uses an outer shield as one of the conductors, so that energy is contained between the outer (shield) and the inner conductors as shown in Figure 52-9.

The waveguide, as the coaxial line, also contains the energy within its outer walls; and prohibits energy from escaping.

Copper loss becomes particularly important at microwave frequencies because of a phenomena known as the SKIN EFFECT. At lower frequencies, as shown in Figure 52-10, the skin effect would be extremely small; but at microwave frequencies it would cause an appreciable loss.

A1. Because there is an impedance from each conductor to ground.

A2. It can be used as an insulator if it is a quarter-wave long at the operating frequency.

A3. As frequency increases, wavelength decreases.

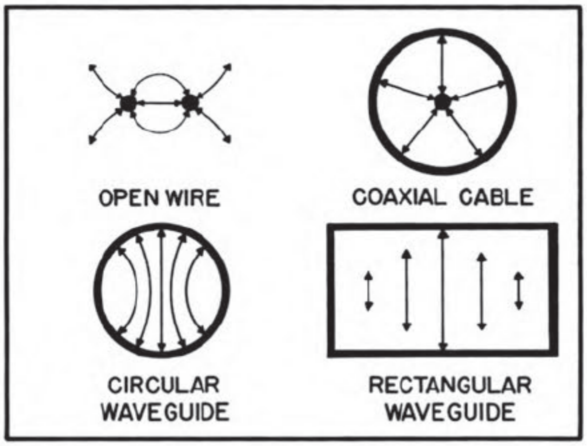


Figure 52-9 - Energy loss in transmission lines.

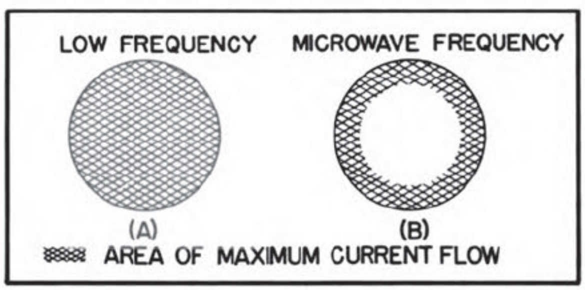


Figure 52-10 - Skin effect.

In a coaxial line, the outer conductor would not have an appreciable resistance; but the small inner conductor would have appreciable resistance because of the skin effect. By removing the inner conductor, the skin effect would be greatly reduced. A coaxial line with no center conductor is a waveguide.

Dielectric loss becomes greater in coaxial lines at higher frequencies. The energy is lost in the insulating material between the inner conductor and the outer shield. Air has a negligible dielectric loss at any frequency. Waveguides have a negligible dielectric loss because they are filled with air or some other gas.

Figure 52-11 shows a comparison between coaxial lines and waveguides. Notice that the waveguides can be designed for a specific range of frequencies. As the frequency changes, a different size waveguide could be used, so that at the new frequency the new waveguide would be operating at its point of minimum attenuation.

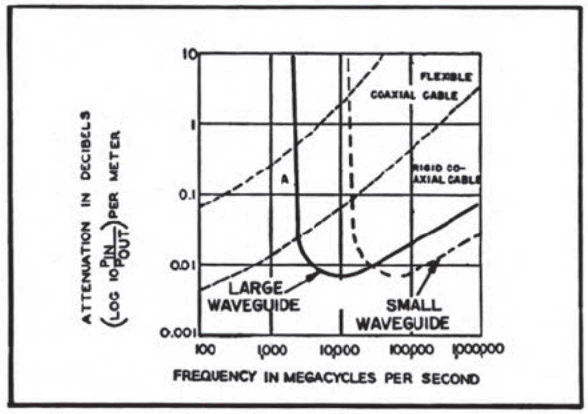


Figure 52-11 - Attenuation comparison.

A second advantage of using waveguides is their increased power handling capabilities as compared to a coaxial line of similar size. The voltage required to break down the insulation would be determined by the distance between the two conductors. Figure 52-12 shows that the waveguide would require much higher break down voltage since the distance is more than doubled. In special applications, the waveguide may be under pressure or filled with gas to extend its power handling capabilities. Another advantage is the waveguide is simple and rugged. Since the waveguide is hollow, there would be no inner conductor which could be damaged by shock or vibration.

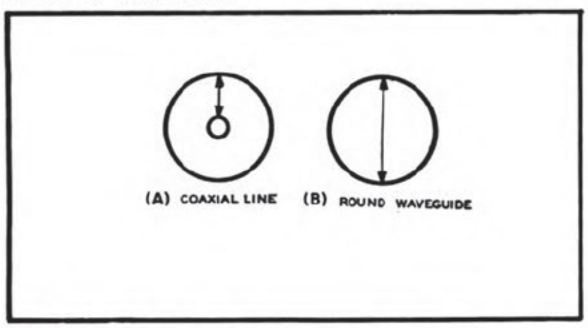


Figure 52-12 - Comparison in breakdown paths.

Q4. Which waveguide dimension determines the power handling capabilities of the waveguide?

Q5. At microwave frequencies, what is the advantage of using waveguides instead of the conventional two-wire line?

## 52-4. Disadvantages of Waveguides

Size is one of the limiting factors in waveguides. One requirement for the transfer of energy through a waveguide is that the width of the wide dimension must be greater than .5 the wavelength of the frequency being transmitted. A frequency lower than this would not propagate through the waveguide. The lowest frequency which will travel through a waveguide is known as the cut-off frequency. It is interesting to note that the frequencies just above the cut-off frequency will have the least amount of attenuation, and is therefore most frequently used. To operate a waveguide at 300 megacycles per second would require a waveguide ("a" dimension) of 50 centimeters. As the frequency decreases, the size of the waveguide must be larger. For example, to transfer 60 cycles per second, the "a" dimension would have to be approximately 1550 miles wide.

The installation and operation of a waveguide transmission system is somewhat more difficult than for other types of line. The radius of bends in the guide must be greater than two wavelengths to avoid excessive attenuation. This requirement may hamper installations in restricted spaces. If the guide is dented, standing waves will be set up: the attenuation offered to a signal traveling through the waveguide is greatly increased. Such faults limit the power handling capacity of the system, and make the possibility of arc-over more likely. Unless great care is exercised in the installation, one or two careless joints may nullify completely the advantage obtained from the use of the waveguide.

Summarizing, the characteristics of waveguides are as follows:

1. Advantages

    a. Less loss
    b. Greater power handling capacity
    c. Simpler construction and ruggedness

2. Disadvantages

    a. Size of the waveguide at lower frequencies
    b. Difficulty of installation

3. Characteristics

    a. The "b" dimension determines the maximum transmittable power
    b. The "a" dimension determines the minimum transmittable frequency (cut-off frequency)

Q6. What is one disadvantage of waveguide?

## 52-5. Boundry Conditions

The output of the radar transmitter is fed to the waveguide. This energy is radiated into the waveguide just as the Marconi and Hertz antennas radiate electromagnetic energy into free space. The only difference between antenna and waveguide propagation is that the waveguide contains and directs the RF energy.

Since this is electromagnetic energy, it is convenient to speak of the energy in terms of electric (E) and magnetic (H) fields. Electromagnetic radiation is a TRANSVERSE WAVE, that is, the wave motion is perpendicular to the direction of propagation. Waves in water are transversial in nature. The wave motion is up and down, and it travels in a horizontal direction. The electric and magnetic fields that exist in the waveguide must do so under specific conditions. These are known as BOUNDARY CONDITIONS.

One condition that must be met by the electromagnetic field within a waveguide is that the field must be continuous throughout the region in which the dielectric is constant. This condition requires that the frequency at one point in the waveguide be the same as the frequency at any other point.

A second and more important condition may be stated as follows: the electric field (E-field) which exists in a waveguide is always perpendicular to the surface on which it acts. The magnetic field (H-field) is in a direction parallel to the surface of the waveguide. The H and E fields are always perpendicular to one another. The relationship of these fields and their relative magnitudes within the waveguide will be fully explained when necessary.

## 52-6. Energy Propagation in a Waveguide

An understanding of wave motion in free space is helpful in understanding how a transverse wave would act when it travels through a waveguide. An expanding wave front, after it has moved several wavelengths from its point of origin, can be considered to be an almost straight wavefront if the considered portion is small. If this wavefront is placed in the confines of a waveguide, the field will not be able to expand at random. There will be a confining action that will cause the wavefront to pass down the guide. The wavefront moves down the guide because of its reflection from the walls of the guide. This action may be seen by referring to the diagram in Figure 52-13. This diagram shows a single wavefront composed of small particles.

Particle 1 strikes the wall, and acting as if it were a rubber ball, is bounded from the wall at approximately the same velocity. By striking the wall, its forward movement is not

A4. The b dimension.

A5. Less radiation loss, less attenuation.

A6. They are bulky at low frequencies. A two wavelength bend is required.

arrested but redirected. If the wall is perfectly flat, the angle of incidence will be the same as the angle of reflection. An instant later, particle 2 will strike the wall and be reflected. Because all the particles are traveling at the same velocity, particles 1 and 2 will not change their relative position with respect to each other; and the reflecting wave will retain the same shape as the original. Particles 3, 4, and 5 will reflect from the wall in the same manner resulting in a new wavefront identical to the original with one exception. The reflected wave will be inverted with respect to the incident wave.

An antenna in free space radiates in all directions. However, in the waveguide only those components striking the surface of the waveguide wall will be considered. After being radiated into the guide, the wave will be reflected many times as it travels toward the end. In Figure 52-14, the top view of an antenna is shown. At point 1 energy traveling to point 2 will be reflected from the wall to point 3. In Figure 52-14, only the E-field is shown with the wavefronts perpendicular to the direction of propagation. The alternations of the wavefront are indicated by the positive and negative signs. The wavefronts, although curved, are represented as a flat plane. The distance between any two repeating voltages, positive or negative, represent a wavelength of the transmitted frequency. After striking the wall, the wavefront will be reflected at an angle equal to the angle of incidence. It is important to remember that the wavefront will be inverted when it is reflected. In other words, a positive wavefront will be reflected as a negative wavefront, and vice versa. Since this is true, at the point of reflection, the E-field will cancel at point 2, and effectively be nonexistent. In Figure 52-15, the incident and reflected waves are shown in their combined form. Along the entire surface of the wall, the resultant E-field will be zero because the incident and reflected waves are out of phase. At the line A-A', the energy will be in phase and will be additive. Continuing outward from the wall to line B-B', the result would be a condition similar to that at the wall. The incident and reflected would be out of phase and effectively no E-field would exist there. Since the line B-B' represents a zero E-field, or POTENTIAL LINE, another metal wall (which itself represents a condition of zero voltage) could be placed along line B-B' with no effect on the wavefronts between them. The second wall could be placed anywhere the incident wavefronts and the reflected wavefronts combine to give a zero potential line such as line B-B', or D-D'. The addition of the second wall forms the "a" dimension.

Each successive zero line in which the second wall is placed will result in a different type, or MODE, of operation. Generally, if the spacing between the walls is the smallest, the waveguide is operating in the DOMINANT mode.

Only the E-field has been shown so that the reflections could easily be understood. The radiated energy is contained in the form of both

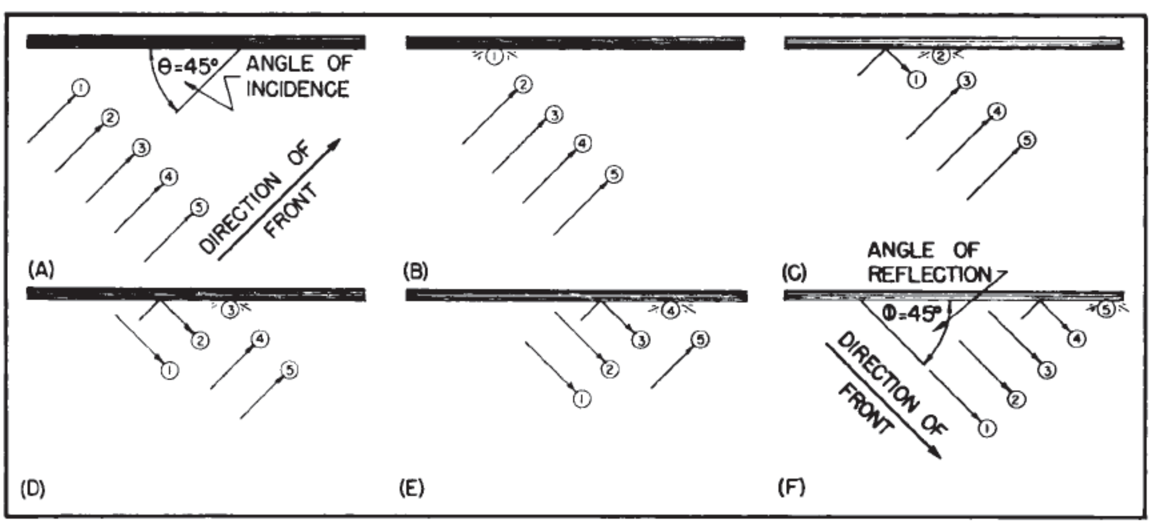


Figure 52-13 - Reflection from a plane surface.

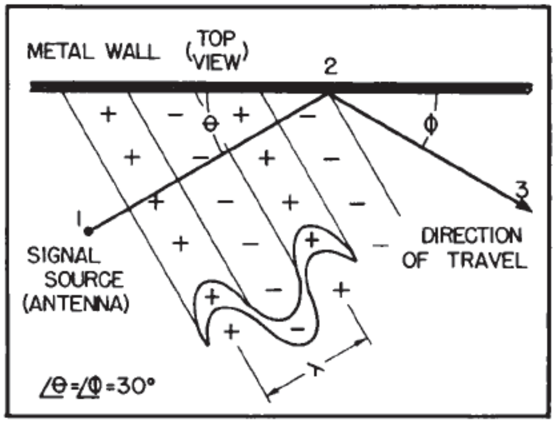


Figure 52-14 - Wavefront incident to a plane surface.

E and H fields. One cannot exist without the other. Since energy is being reflected from both walls, they will combine and form a resultant field. In the process of radiation, an H-field will be generated as shown in Figure 52-16B. The H-field will always exist in closed loops. This field is the resultant H-field. The combinations of both the E- and H-fields will be the resultant electromagnetic energy which will, as shown in Figure 52-16C, propagate down the guide.

NOTE: The length of the E lines in Figure 52-16A indicate the intensity of the field and not the starting and terminating point of the field.

A simple rule used to establish the relationship of the direction of travel to the electric and magnetic fields is known as the POYNTING VECTOR. It states that a screw (right hand thread) with its axis perpendicular to the electric and magnetic fields, would advance in the direction of propagation if the E-field vector were rotated toward the H-field vector through the smallest angle. This is sometimes referred to as the Right Hand Rule for Electromagnetic Energy. A diagram showing how the relationship of the fields to the direction of travel is illustrated in Figure 52-17.

The cut-off frequency ($f_{co}$) of a waveguide is the lowest frequency that will propagate through the guide. It has been stated that energy will radiate in all directions from the antenna or probe. Since only a small portion of the radiated energy will travel straight down the guide, (this will happen above or below the cut-off frequency) the energy that will be reflected from the walls of the guide will be the only portion considered. The area of maximum voltage as shown in Figure 52-18, will occur in the center of the guide; and the miminum voltage must occur at the wall of the guide. Because of this fact, a definite pattern will be formed inside the guide. The energy contained near the walls will be in the form of a magnetic field.

By examining Figure 52-19, it can be seen that only at one angle will the reflections add to create a maximum E field in the center of the guide, and not at the sides. Notice in Figure 52-19B that the line 0-0' intersects two zero points that are at directly opposite points on the wall. In this condition, energy can exist; and will propagate through the waveguide. At any other angle, a resultant E field pattern would be developed that would require an E field to exist at the side of the wall as shown in Figure 52-19 A, C.

In sections of waveguide in which the "a" (wide) dimension is held constant, varying the frequency of the transmitter will cause the path of propagation to change. This can be seen by

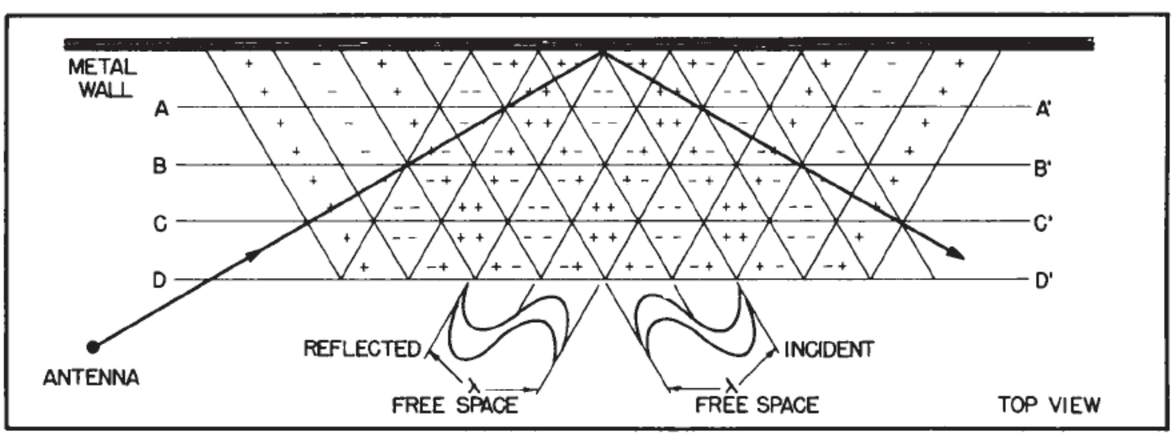


Figure 52-15 - Combination of wavefronts incident to, and reflected from, a plane surface.

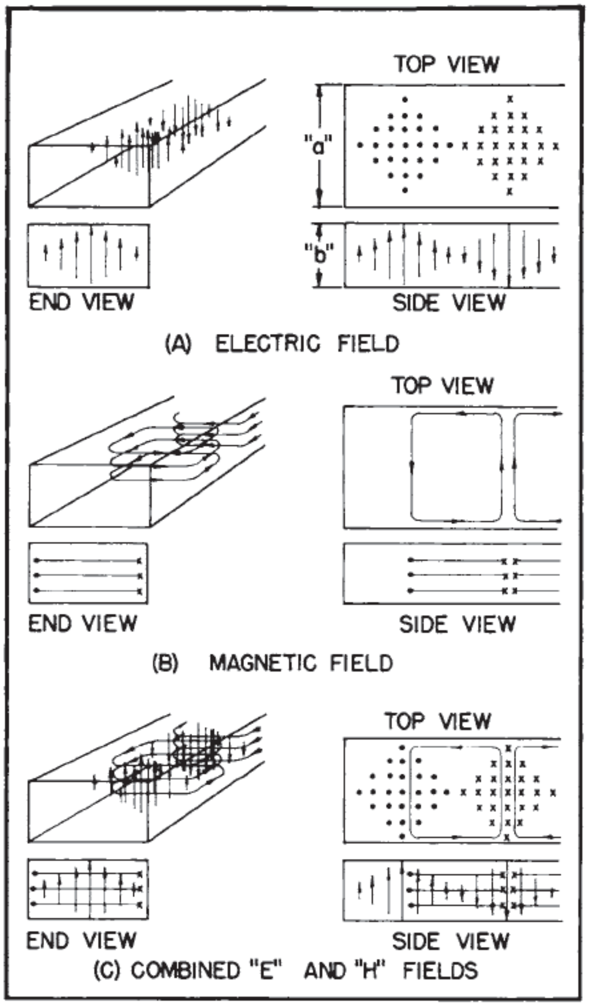


Figure 52-16 - Electric and magnetic fields in a waveguide.

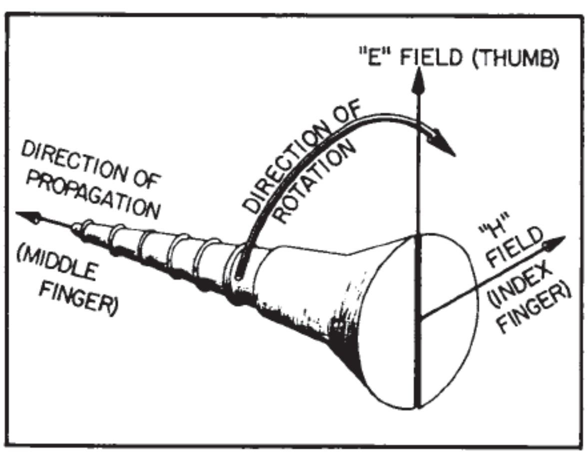


Figure 52-17 - The Poynting vector.

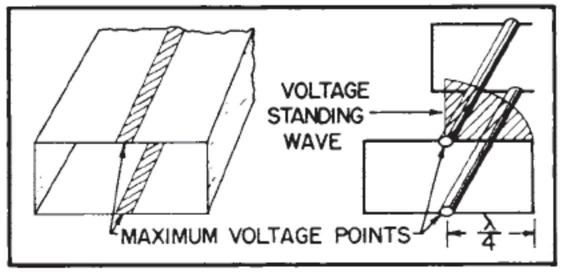


Figure 52-18 - Area of maximum voltage.

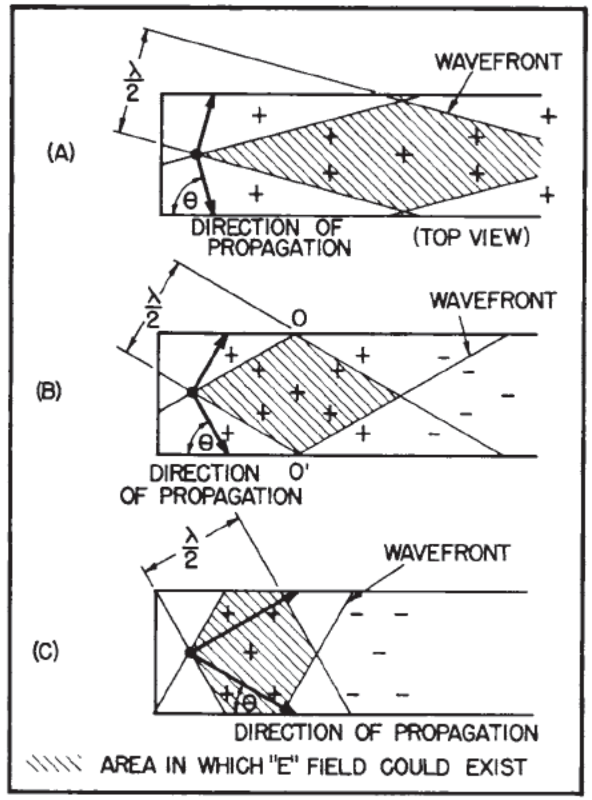


Figure 52-19 - Effect of the "E" field on the angle of propagation.

applying the rule which states that the zero point occurs at directly opposite points on the wall. This is illustrated in Figure 52-20. As the frequency is made lower and approaches the cut-off frequency, the direction of travel becomes more perpendicular to the side of the wall. At the cut-off frequency, the wavefront will travel back and forth across the guide with no forward motion.

It should be noted that a wavelength in the guide appears to be longer than in free space. Due to the reflecting wavefronts combining at an angle as shown in Figure 52-21, the wavelength inside the guide will always be longer than in

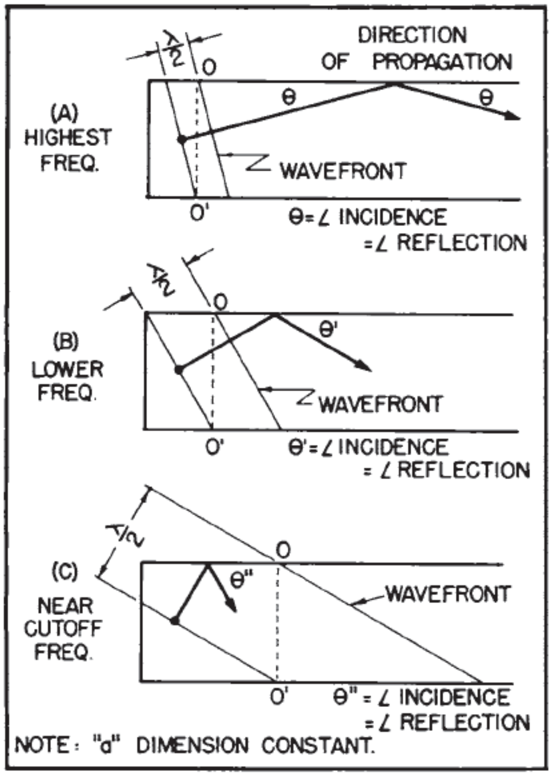


Figure 52-20 - Effect of frequency on the angle of propagation.

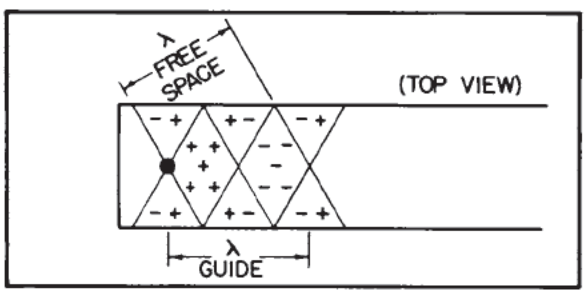


Figure 52-21 - Comparison of free space wavelength to guide wavelength.

free space; and is referred to as the GROUP WAVELENGTH ($\lambda$ g). This is important in determining bends, twists and the location of coupling devices.

The velocity of RF energy is considered to be 186,000 miles per second. This will be referred to as FREE SPACE VELOCITY ($V_o$). This is shown in Figure 52-22.

An interesting phenomena occurs in the waveguide when the wavefront strikes the wall. In Figure 52-23, the wavefront is shown moving

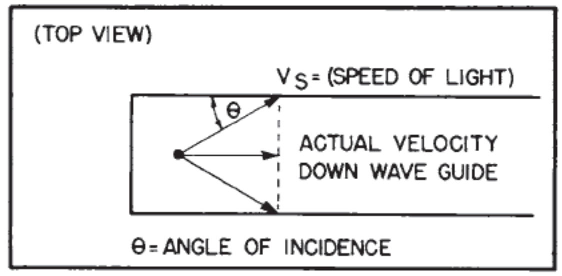


Figure 52-22 - Relationship of free space velocity to group velocity.

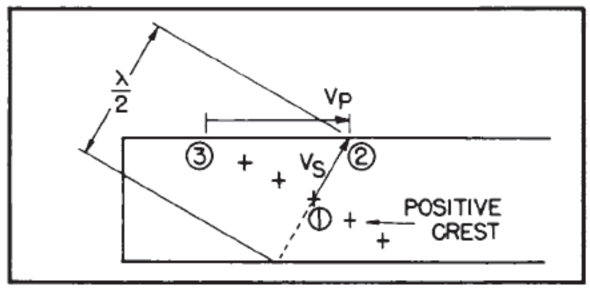


Figure 52-23 - Comparison of free space velocity to phase velocity.

toward point 2 (at the speed of light). In the time it takes, the positive crest will also travel along the wall from 3 to 2 in the same time. Since the distance is greater, its speed will also be greater. Depending on the angle of incidence, it can be several times the speed of light. This is known as the PHASE VELOCITY ($V_p$).

Q7. What is the relationship between the E and H fields in a waveguide?

Q8. What is the relationship between the angle of incidence and the angle of reflection?

Q9. What is the relationship of the propagated E fields and the walls of the waveguide?

Q10. What is the dominant mode?

52-7. Modes of Operation

Thus far, only the most basic type of E and H field arrangement has been shown. It is sometimes necessary to have a more complex arrangement to facilitate coupling, isolation, or types of operation. To describe the various arrangements or Modes of operation, the field arrangements are first divided into two categories: TRANSVERSE ELECTRIC and TRANSVERSE MAGNETIC.

A field is transverse electric (TE) when the electric field is perpendicular to the sides of

A7. They are $90^{o}$ out of phase.

A8. They are equal.

A9. The "E" field components are perpendicular to the walls and are propagated in a direction parallel to the walls.

A10. The mode which was the lowest cutoff frequency.

the guide and has no components along the lengthwise axis of the guide.

A field is transverse magnetic (TM) when the magnetic field has no components along the length-wise axis of the guide.

Secondly, subscripts are given to designate the exact field arrangement. In rectangular waveguides, the first subscript following the TE or TM designation indicates the number of half wavelengths in the "A" or wide dimension as shown in Figure 52-24. The second subscript indicates the number of half wavelengths in the "B" dimension. For circular waveguides, the first subscript indicates the number of whole wavelengths around the circumference. The second subscript indicates the number of half wavelengths along the diameter.

The mode of operation most often used is the DOMINANT mode. For a given size of waveguide, this is the mode that has the lowest cut-off frequency. For rectangular waveguides the $TE_{1,0}$ would be the dominant mode, and for circular waveguides it would be $TE_{1,1}$. The dominant mode has the advantage of the least attenuation.

In circular waveguides, the $TM_{0,1}$ is the most often used due to its ease of termination. Because of its symmetry, the circular waveguide is especially useful with rotating sections, where one section is able to rotate without causing a change in its ability to transfer energy. Some of the more common modes are shown in Figure 52-24.

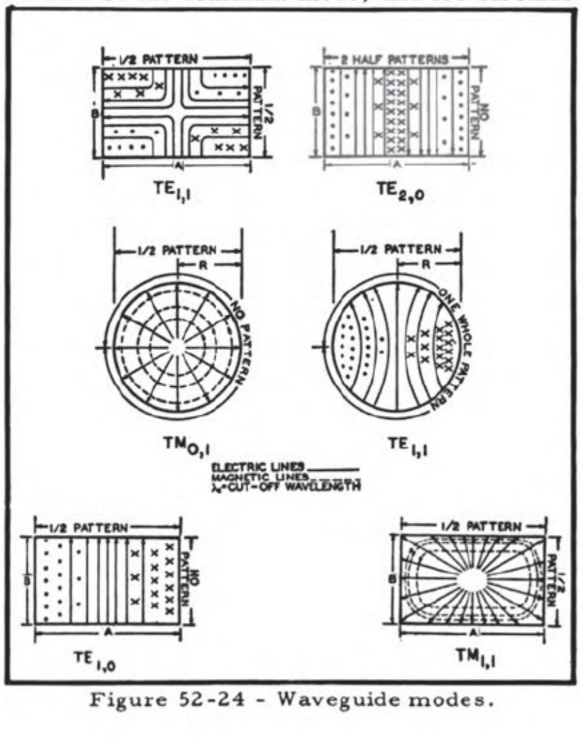


Figure 52-24 - Waveguide modes.

Q11. What is a transverse wave?

Q12. What is the advantage of the dominant mode?

## 52-8. Methods of Coupling and Impedance Matching

There are basically three ways to couple energy into and out of a waveguide. Each has its own characteristics, and the choice of which one is used is based on the external physical limitations and design considerations (size, mode of operation, attenuation, impedance matching, etc.). The coupling between waveguides and coaxial lines will be discussed first.

The probe, or capacitive coupling, is shown in Figure 52-25. Its action is the same as a quarter-wave antenna. The outer conductor (shield) is connected to the waveguide proper, and the inner conductor is inserted into the guide. The probe should be placed in the center of the "a" dimension and a group of wavelengths from the closed end of the guide. Wide-band probes are often used so that there will be no serious

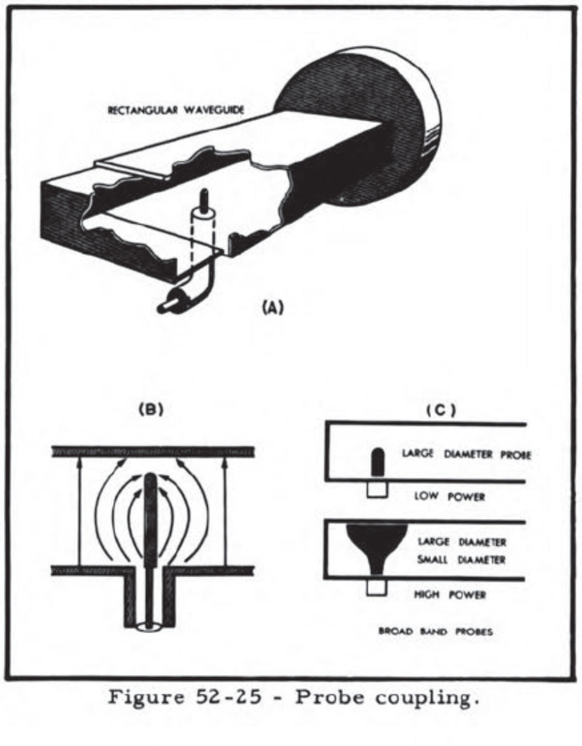


Figure 52-25 - Probe coupling.

attenuation of a pulsed radar's associated side-bands.

Figure 52-26 shows LOOP or inductive coupling. The loop may be placed anywhere on the guide as long as it can be cut by the magnetic field. Usually it is placed at a location where the H lines are maximum as in Figure 52-26B. The outer conductor of the coaxial line is grounded to the waveguide chassis, and the inner conductor forms a loop inside the guide. It is then grounded. The large current that flows establishes a correspondingly large magnetic field, and transformer action transfers the energy.

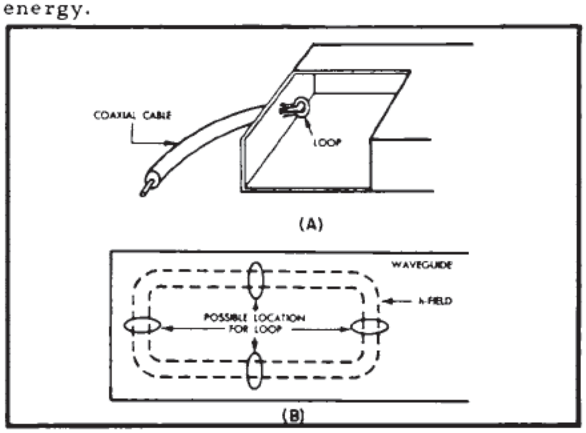


Figure 52-26 - Loop coupling.

The third method usually employed as a coupling device between two cavity sections, or between a resonant cavity and waveguide, is called APERATURE or SLOT coupling. This type of coupling is shown in Figure 52-27. The idea here is that if a hole is put in a waveguide wall, some of the energy will leak through. Depending on the location of the slot, the coupling can be E lines, H lines or E and H lines combined. In Figure 52-27, slot A shows E line coupling, slot B shows H line coupling, and slot C shows E and H line, or electromagnetic coupling. As a rule, if the hole breaks a current path, or if the hole disrupts the electric field where it is not equal to the H field, the transfer of energy will occur. Any one of these methods mentioned can be used to couple energy into or out of the waveguide.

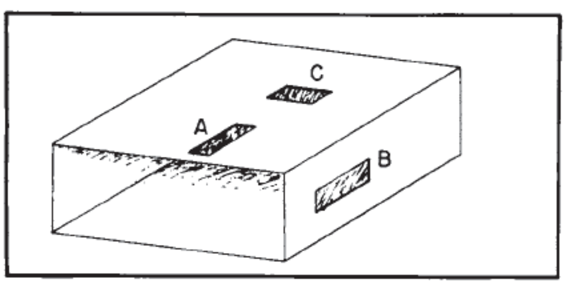


Figure 52-27 - Aperature or slot coupling.

When connecting two waveguides of different size and different impedances, several methods may be employed. A section of waveguide called a FLARED MATCHING SECTION is used to match impedances. It is a device which has a gradually expanding dimension. It is shown in Figure 52-28. Flaring permits the electromagnetic fields to expand or contract smoothly to fit the new size of waveguide. If the change is made gradually, there will be no reflections back into the original waveguide section.

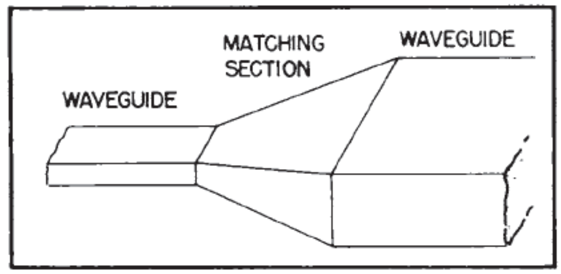


Figure 52-28 - Flared matching section.

This same principle is used to match the impedance of the waveguide to free space. The FEED HORN permits the field to expand during transmit, or contract during receive. The flaring is gradual to prevent the formation of standing waves. The open end of the feed horn, shown in Figure 52-29, is covered with a dielectric material (usually plastic) to prevent foreign material from entering into the guide resulting in reflection.

Quarter-wave sections of waveguide can also be used to match impedances between different size waveguides. This matching section is shown in Figure 52-30. The size of the matching section and the size of its apertures can be made to give a proper impedance match.

If the connection is to be made between two waveguides of the same size, a choke joint is used. A typical choke joint is shown in Figure 52-31. Solder joints would not be practical because if solder were permitted to drop inside the joint, reflections would occur. Also if the

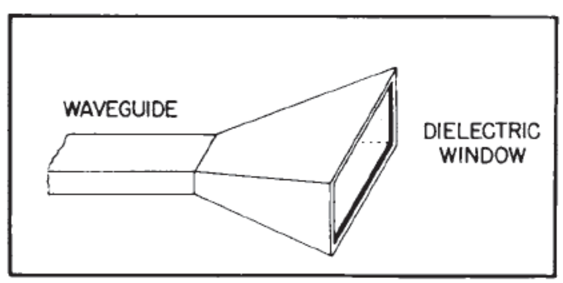


Figure 52-29 - Feed horn.

A11. The wave is transverse when its motion is perpendicular to the direction of propagation.

A12. It possesses the least amount of attenuation and is the most efficient mode.

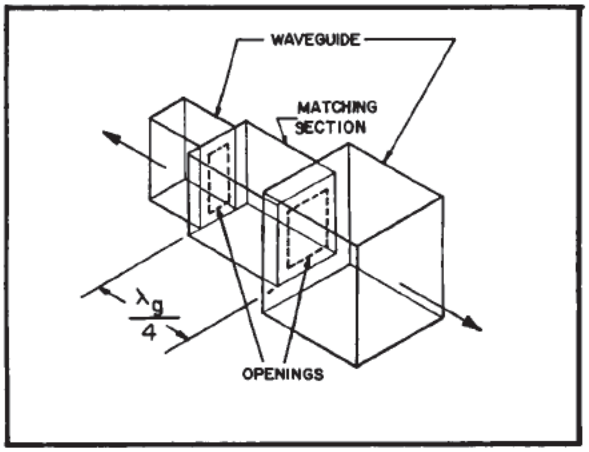


Figure 52-30 - Quarter-wave matching section.

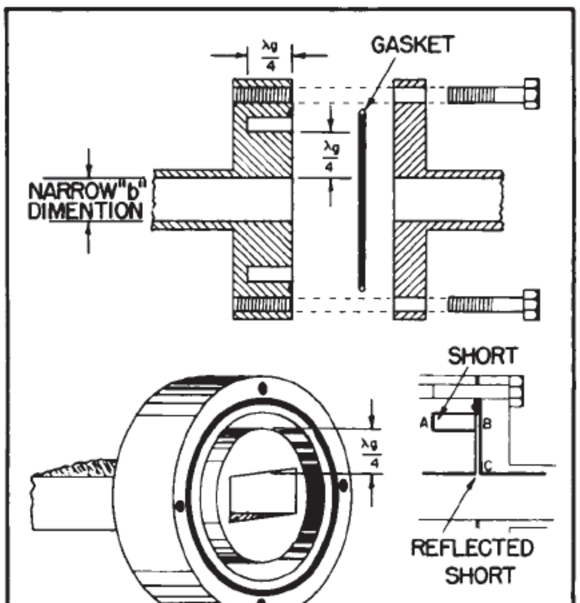


Figure 52-31 - Choke joint.

joint was not completely sealed, energy would be lost. By using the choke joint, both of these disadvantages are overcome. The distance from A to B is a quarter wavelength, and the distance from B to C is also a quarter wavelength. The short circuit at point A will reflect back a short to the inner surfaces of the wall. Since this is the normal condition at the surface of the wall, there will be little or no energy lost at the joint; and the energy will travel past with no effect on the field patterns.

Very seldom will a radar set be found which has no bends or twists in the waveguide system. Since any abrupt change in the waveguide size would cause reflections, all bends and twists should be made gradually. Figure 52-32 shows the more common bends and twists. If the bends have a radius of at least two wavelengths, the energy will be able to complete the bend with a minimum of attenuation. The twists will be made so that it will take place over a distance of greater than two wavelengths so that there is little attenuation.

If a small mismatch occurs in a waveguide system, an inductive or capacitive impedance can be introduced near the mismatch, which if of the proper value, will make the waveguide act as a non-resonant or matched system. METAL PARTITIONS or IRISES are used as matching devices.

The metal partitions are placed in the guide as shown in Figure 52-33. The inductive partition is placed in the area of maximum H lines. The partition will change the permisability which will, in turn, change the value of inductance and impedance. This action is similar to inductive tuning of a tank circuit by varying the permisability of the inductive tuning coil.

The capacitive partition is located in the area of maximum E lines. The capacitive partition will change the distance across which the E lines must exist, which will change the capacitance and the impedance. This is similar to changing the distance between the plates of a capacitor to vary the capacitor. One disadvantage of using a capacitive partition, however, is the

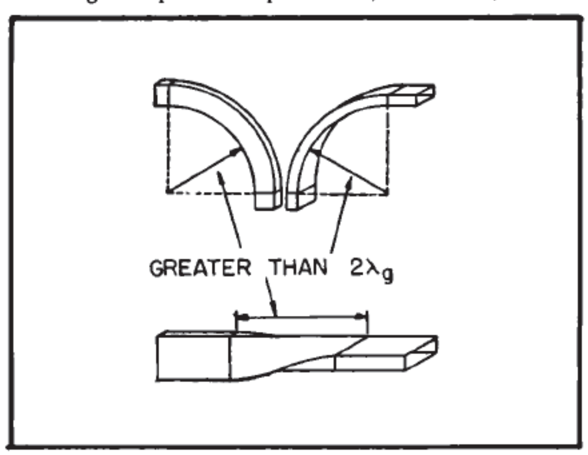


Figure 52-32 - Waveguide bends and twists.

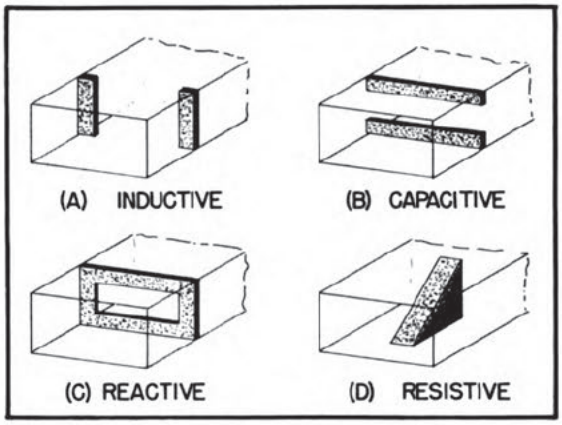


Figure 52-33 - Waveguide partitions.

lowering of the power handling capabilities because the "b" dimension is made smaller.

If the partition is located in both the area of E and H fields, the partition is considered reactive. Depending on its size, it could be capacitive, inductive, or resistive. A reactive partition may appear as a high resistance to the dominant mode, but to a higher mode it could appear as a low shunt impedance. In this manner, it could be used very effectively as a mode suppressor. It would have no effect on the desired or dominant mode.

## 52-9. Terminating a Waveguide

Since a waveguide is a single conductor, it is not as easy to define its characteristic impedance ($Z_o$) as it is for the coaxial line. It can, however, be thought of as being approximately equal to the ratio of the strength of the electric field to the strength of the magnetic field for energy travelling in one direction. This ratio is equivalent to the voltage to current ratio in coaxial lines on which there are no standing waves.

Four methods which are used to terminate waveguides are shown in Figure 52-34. In Figure 52-34A, the energy is absorbed by a flat plate of resistive material placed across the inside of the guide. In Figure 52-34B, the energy is dissipated by a rod of resistive material across the guide. In Figure 52-34C, the energy is dissipated by a small particle of resistive material suspended across the high voltage points. In Figure 52-34D, the energy is dissipated by a mass of carbon-coated cloth or graphite sand.

In general, the resistive material acts as a resistive load to dissipate energy through $I^2R$ losses. If the material has a low impedance, it should be placed where the ratio of E lines to H lines is low; however, if the material has a

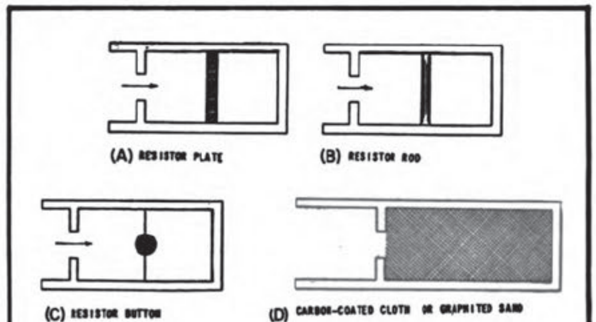


Figure 52-34 - Methods of terminating a waveguide.

high impedance, it should be placed where the E lines are maximum. This condition is similar to picking the proper E/I ratio on a two-wire resonant line for matching purposes.

A resistive partition is placed in the guide so that it will dissipate energy. It is usually some carbon or graphite impregnated material. Its purpose is to serve as a resistive load to minimize reflections. It should be realized that by altering the size of a waveguide will change its characteristic impedance. The partitions are just one of the many ways of changing impedances.

Q13. What are three types of coupling?

Q14. Why must a waveguide be flared?

Q15. What is the purpose of a partition?

## 52-10. Cavity Resonators

The first type of resonant circuit encountered in ordinary radio work usually consists of a coil of wire with a capacitor shunted across its terminals. This is shown in Figure 52-35A. To increase the resonant frequency, the inductance, capacitance or both must be reduced, since the resonant frequency equals:

$$f_o = \frac{1}{2\pi\sqrt{LC}} \qquad (11\text{-}16)$$

A frequency will finally be reached at which the inductor, L, is a single turn of wire, and the capacitor, C, consists of the distributed capacitance across the opposite sides of the same turn.

The resonant frequency cannot be increased further by the addition of several quarter-wave (Lecher Line) lines in parallel as shown in Figure 52-35B, because connecting the lines in parallel decreases the inductance in the same proportion that the capacitance is increased leaving the resonant frequency unchanged. How-

A13. Loop, slot, and probe coupling.

A14. To prevent standing waves.

A15. Their use will change the "a" and "b" dimensions while accomplishing a correct impedance match.

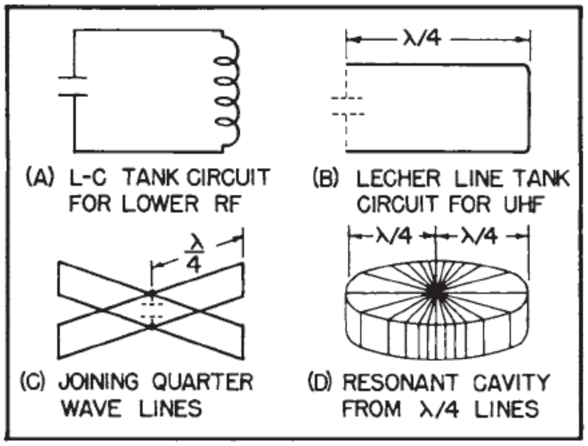


Figure 52-35 - Development of cavity resonators.

ever, an important benefit is gained by paralleling the quarter-wave lines. The area in which current can flow is greatly increased, thereby decreasing the resistance, and increasing the circuit Q.

Since the resonant frequency is not affected by the number of quarter wavelines connected together at their open ends (high impedance points), the diagram of Figure 52-35C can be filled in completely to form a cylinder with closed ends as shown in Figure 52-35D. This cylinder is called a CAVITY RESONATOR. The function of a cavity resonator is similar to that of any coil and capacitor resonant circuit. The Q is high, and the circuit is very selective. It will only resonate over an extremely narrow frequency range. Because of its physical size, the cavity resonator is practical only in the microwave frequency range. For example, at one megacycle the cavity would be about 100 feet long; but at 1000 megacycles, the size would be measured in inches instead of feet.

The boundry conditions for waveguides also apply to cavity resonators because the fields generated are entirely inside the cavity as they are in a waveguide. No E or H lines exist outside the closed cavity, and electron flow is limited to a thin layer of metal on the inside surface of the cavity.

In the cavity, the electric and magnetic fields are 90° out of phase which will produce standing waves. A comparison can be made between a flat transmission line and a resonant transmission line where the resonant line has the voltage and current standing waves 90° out of phase.

Modes in cavities usually are designated by a three-number system instead of the two-number system used in waveguides. The third subscript signified the number of half-wave patterns crossed perpendicular to the transverse field. Thus, the mode of Figure 52-36 is classified as TM 0, 1, 0 for a cylindrical cavity.

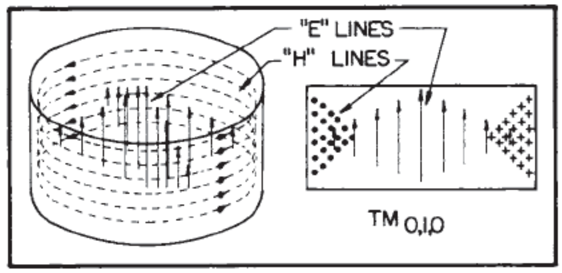


Figure 52-36 - Fields in a cavity resonator.

The same cavity can oscillate at several different modes depending upon the manner in which it is energized. Standing waves may be formed by reflection from various surfaces so that a cavity usually can resonate at several fundamental frequencies as well as harmonics of the fundamentals. Figure 52-37 shows cavity resonators of various shapes.

The flow of electrons in a cavity resonator is confined to an exceedingly thin layer of metal on the inside surface of the cavity. This layer

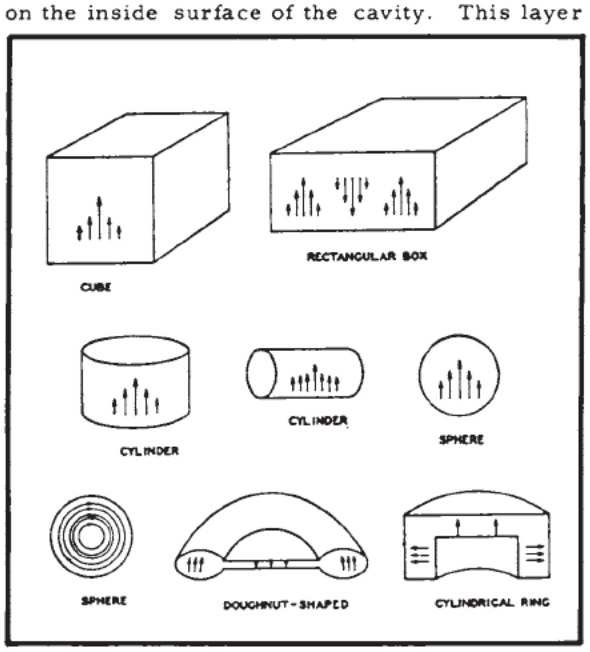


Figure 52-37 - Cavity shapes.

should present as low a resistance as possible so that loss will be negligible. Cavities may be made from thin sheets of copper or from some metals plated with silver, copper, or gold on the inside. In some cases, cavities may be constructed from nonconducting materials with the inside sprayed with a thin layer of metal or covered with metal foil. Cavities used for precision frequency measurement often are made by hollowing out solid blocks of metal so that the dimensions (and resonant frequency) do not change. The extra amount of metal is used for mechanical strength and rigidity only, and contributes nothing towards lessening the resistance to the high-frequency current flow.

Energy is coupled into and out of cavities by the same methods used in waveguides i.e. probe, loop and aperture, or slots.

There is an additional way to couple energy into a cavity. The REENTRANT CAVITY is slightly different than other types of cavities. An illustration of the reentrant cavity is shown in Figure 52-38. The center portion has openings through which a stream of electrons are passed. If the electrons are formed in bunches or groups, the edges of the center wall will feel a varying potential as the bunches of electrons travel through. The dense bunch of electrons are negative with respect to a less dense bunch. As the electron bunches pass through the cavity, if properly timed and spaced, they will give up energy to the fields in the cavity. This method is commonly employed in microwave oscillators.

The Q of the reentrant cavity is lower than other cavity shapes (approximately 4000 as compared to cylindrical type which may be as high as 31,000).

Energy can be coupled out of a reentrant cavity in the same manner as it was coupled in. An electron beam travelling through the cavity will form in bunches as it is passed through the cavity if there is an existing oscillating field.

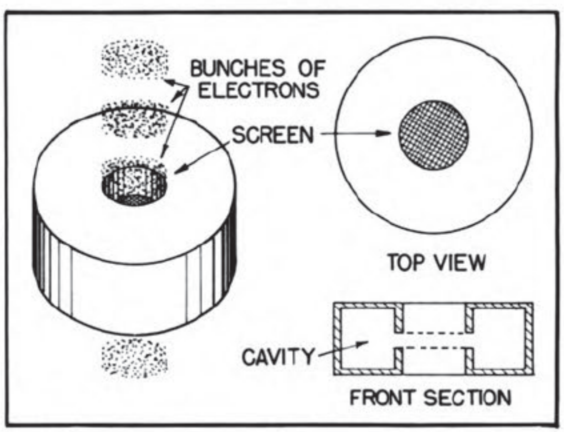


Figure 52-38 - Reentrant cavity.

This forming or bunching is taking energy away from the existing field.

The resonant cavity which is used as a tank circuit at microwave frequencies may often require tuning. Tuning is accomplished by changing the size of the cavity to achieve a new resonant frequency. The first method will be changing the size or volume of the cavity as shown in Figure 52-39A. By moving the adjustable plunger, the half-wave dimension is changed. This would cause the cavity to have a half-wave dimension for another frequency. If the volume is decreased, the cavity will resonate at a higher frequency. If the volume is increased, the resonant frequency would be lower.

Figure 52-39B shows capacitive tuning. An adjustable slug or screw is placed in the area of maximum E lines. D1 represents the distance between two capacitor plates. As the slug is

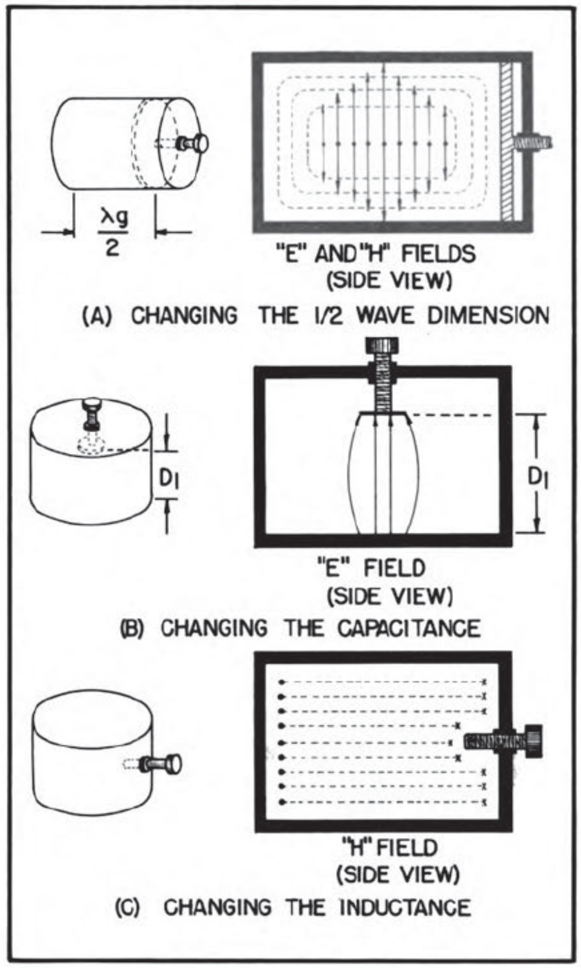


Figure 52-39 - Methods of varying the resonant frequency of the cavity.

moved in, the distance between the two plates would become smaller, and the capacitance would increase causing a decrease in the resonant frequency of the cavity. If the slug is moved out, the resonant frequency of the cavity would increase.

By placing a non-magnetic slug in the area of maximum H lines (Figure 52-39C), a similar effect is obtained. By moving the inductive slug in, the magnetic lines of flux are interfered with. The further into the cavity the slug is moved, the more magnetic lines of flux it will interfere with. This decreases the permisability of the space in which flux can exist, lowering the inductance, and raising the resonant frequency of the cavity. By moving the slug out, the resonant frequency will decrease.

In most applications a micrometer type adjustment is used for tuning to insure some amount of accuracy. Some of the main uses of resonant cavities are: impedance matching, microwave oscillators, filters, and wavelength or frequency measurement.

Q16. How do the mode designations for the cavity resonator differ from those used in waveguides?

Q17. What are the three methods of cavity tuning?

## EXERCISE 52

1. What is the relative magnitude of the input impedance of a shorted quarter-wave transmission line stub support?
2. How is the waveguide concept derived from the two-wire transmission line supported by quarter-wave stubs?
3. In the course of the evolution from a two-wire transmission line to a waveguide, do the electromagnetic field configurations remain the same? Why?
4. In what form is energy transmitted through a waveguide?
5. What is the relative magnitude of the electric field component of an electromagnetic wave at the boundary surface of a waveguide if the electric field is parallel to the surface?
6. When an electric field (or voltage) attempts to exist across a perfect conductor, what action occurs?
7. What is the relative magnitude of the magnetic field component of an electromagnetic wave at the boundary surface of a waveguide if the magnetic field is perpendicular to the surface?
8. When a magnetic field tends to cut through a conductor and a voltage is induced that sets up a current in the conductor, what action occurs between the magnetic field created by the current and the exciting magnetic field?
9. When a wave is reflected from a conducting boundary, what action occurs?
10. Which velocity is used to compute the length of time required for a wave to propagate through a waveguide of given length?
11. What is the guide width in terms of the wavelength corresponding to the cut-off frequency of the guide?
12. For a given waveguide, is the cut-off frequency the highest or the lowest frequency that can be propagated down the guide without excessive losses?
13. At frequencies slightly above cut-off, is the principal loss due to the skin effect? At higher frequencies, what is the principal loss due to?
14. What is a fundamental law describing the configuration of magnetic flux lines?
15. What is the assumed direction of motion of the field when applying the left-hand rule for induced current (electron flow) in the walls of a waveguide?
16. In designating the mode of operation of a rectangular waveguide, what is the meaning of the letters TE?
17. What does the first subscript number in $TE_{1,0}$ indicate? What does the second subscript indicate?
18. What is the meaning of the designation $TM_{1,1}$ when applied to rectangular waveguides?
19. What is the meaning of dominant mode when applied to a particular waveguide?
20. What is the designation for a dominant mode of operation for rectangular waveguides?
21. State five reasons why the dominant mode for rectangular waveguides in the mode most commonly used.
22. At radar frequencies, why is it impossible to use lumped components of inductance and capacitance?
23. What is the distance from the closed end of a waveguide to the excitation probe in terms of quarter wavelengths?
24. Loop coupling between the coaxial line and cavity resonator couples to what component of the electromagnetic wave?
25. When connecting waveguides of different sizes and shapes together, what two methods of avoiding reflections are employed?

A16. By an additional number. The number signifies the number of half-wave patterns crossed perpendicular to the transverse field.

A17. Changing the physical dimensions of the cavity, and by changing the inductance or capacitance of the cavity by means of a probe.

# CHAPTER 53

# MAGNETRON, DUPLEXERS, AND ANTENNAS

Before the days of modern radar, equipment operated at considerably lower frequencies. Early radar used special triodes as high power oscillators. Triodes have been used as oscillators in radio transmitters for many years and are simple and effective. However, triodes are not efficient at the higher frequencies and the high power levels required for a pulse radar. At the frequencies used presently by most radars, the effect of interelectrode capacitance in the triode becomes very significant. From the formula

$$f_o = \frac{1}{2\pi\sqrt{LC}} \qquad (11\text{-}16)$$

it can be seen that the frequency generated by a tuned circuit increases as the value of L and C decrease. The interelectrode capacitance of a triode can be reduced by making the tube elements smaller or by increasing the separation between the elements. However, as the elements are moved farther apart, electron transit time is increased, an undesirable condition at radar frequencies. Since the interelectrode capacitance can be reduced by increasing the distance between the elements, it appears that the transit time is the primary reason a conventional triode cannot be used at frequencies about 500 megacycles.

Exhaustive research was conducted in an attempt to make the triode serve as an efficient oscillator at radar frequencies but without success. However, special triodes with short, large diameter leads that are known as acorn tubes and door knob tubes were a valuable byproduct of this effort. These tubes are presently used effectively as amplifiers at very high frequencies (VHF) and ultrahigh frequencies (UHF). They are limited in frequency and power handling capacity. They have never been efficient as radar oscillators operating at frequencies greater than 2,000 megacycles.

The invention of the magnetron, by the British, ushered in a new era in radar. Here was a lightweight, high frequency oscillator that could produce a usable output with a power level that was unheard of before. The magnetron was found capable of producing output signals of up to 100,000 megacycles and power levels up to 6 megawatts. Of course, with this discovery came new problems. New circuits had to be designed to operate magnetrons, new types of modulators, new types of filament supply circuits, etc.

## MAGNETRONS

### 53-1. Characteristics and Construction

The magnetron is an oscillator unlike any other that has previously been discussed in this text. The magnetron is a self-contained unit. That is, it produces an ultra-high frequency output within its enclosure without the use of external components such as a crystal, inductors, capacitors, etc.

Basically, the magnetron is a diode and has no grid. A magnetic field in the space between the plate and the cathode serves as a grid. The plate of a magnetron does not have the same physical appearance as the plate of an ordinary vacuum tube. Since conventional LC networks become impractical at radar frequencies, the plate is fabricated into a cylindrical copper block containing resonant cavities which serve as tuned circuits. The magnetron base differs greatly from the conventional base. It has short, large diameter leads that are carefully sealed into the tube and shielded, as shown in Figure 53-1.

The cathode and filament are at the center of the tube. It is supported by the filament leads which are large and rigid enough to keep the cathode and filament structure fixed in position. The output lead is usually a probe or loop extending into one of the tuned cavities and coupled into a waveguide or coaxial line. The plate structure, as shown in Figure 53-2, is a solid block of copper. The cylindrical holes around its circumference are resonant cavities. A narrow slot runs from each cavity into the central portion of the tube and divides the inner structure into as many segments as there are cavities. Alternate segments are strapped together to put the cavities in parallel with regard to the output. These cavities control the output frequency. The straps are circular metal bands that are placed across the top of the block at the entrance slots to the cavities. Since the cathode must operate at high power, it must be fairly

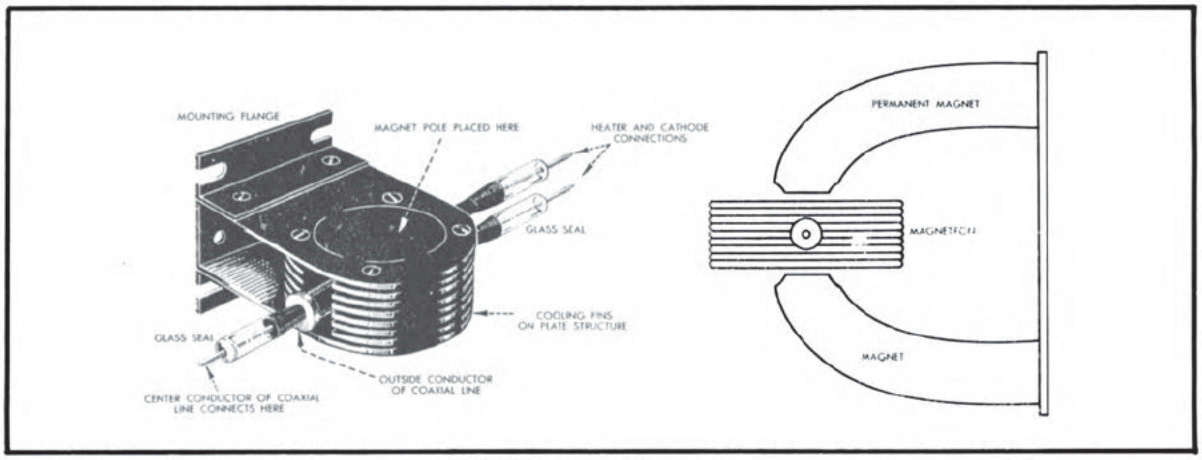


Figure 53-1 - Magnetron.

large and must be able to withstand high operating temperatures. It must also have good emission characteristics, particularly under back bombardment, because much of the output power is derived from the large number of electrons emitted when high velocity electrons return to strike the cathode. The cathode is indirectly heated, and is constructed of a high emission material. The open space between the plate and the cathode is called the INTERACTION SPACE because it is in this space that the electric and magnetic fields interact to exert force upon the electrons.

The magnetic field is usually provided by a strong permanent magnet mounted around the magnetron so that the magnetic field is parallel with the axis of the cathode. The cathode is mounted in the center of the interaction space.

Q1. What controls the output frequency of the magnetron?

Q2. Since there is no grid in the magnetron, what controls the flow of plate current?

## 53-2. Basic Magnetron Operation

The theory of operation of the magnetron is based on the motion of electrons under the influence of combined electric and magnetic fields. The following laws govern this motion.

The direction of an electric field is from the positive electrode to the negative electrode. The law governing the motion of an electron in an electric, or E field, states that the force exerted by an electric field on an electron is proportional to the strength of the field. Electrons tend to move from a point of negative potential toward a positive potential as shown in Figure 53-3. In other words, electrons tend to move against the E field.

When an electron is being accelerated by an E field, as shown in Figure 53-3, energy is taken from the field by the electrons.

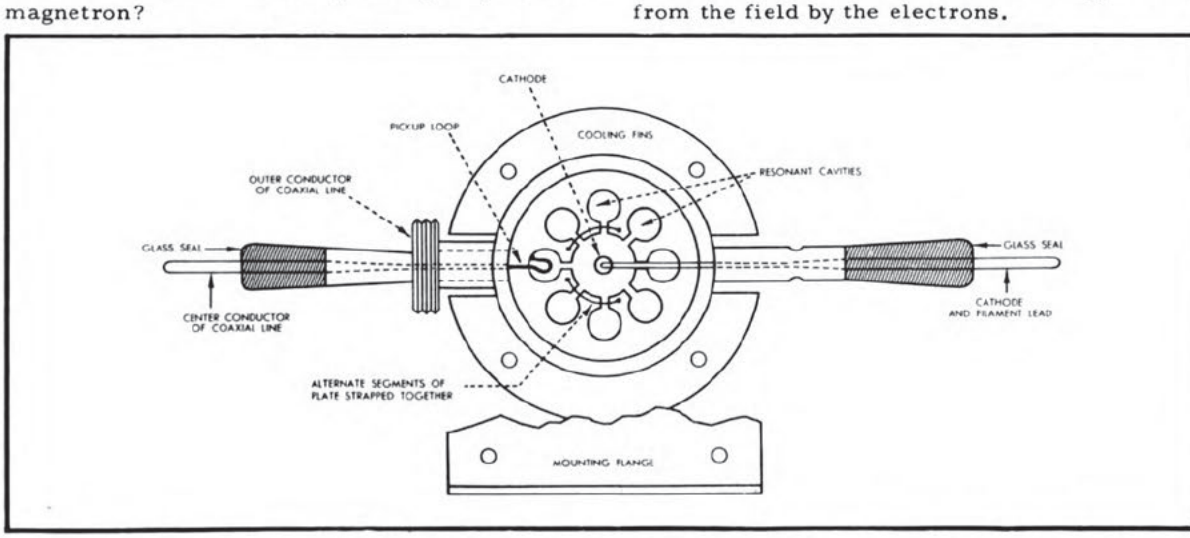


Figure 53-2 - Cutaway view of a magnetron.

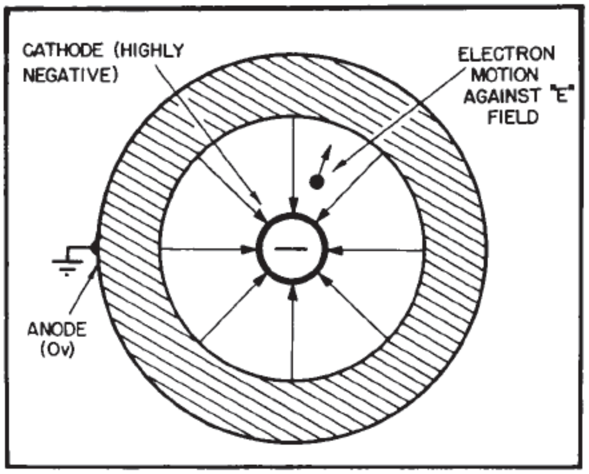


Figure 53-3 - Electron motion in an electric field.

The law of motion of an electron in a magnetic, or H field, states that the force exerted on an electron in a magnetic field is at right angles to both the field and the path of the electron. The direction of the force is such that the electron trajectories are clockwise when viewed in the direction of the magnetic field as shown in Figure 53-4.

In Figure 53-4, it is assumed that a south pole is below the paper and a north pole is above the paper so that the magnetic field is going into the paper. When an electron is moving in space a magnetic field is built around the electron just as there would be a magnetic field around a wire when electrons are flowing through a wire. The familiar left-hand rule can be used to find the direction of the magnetic field existing around a moving electron. Note in Figure 53-4 that the magnetic field around the moving electron adds to the permanent magnetic field on the left side of the electron path and subtracts from the permanent magnetic field on the right side of the electron path, thus weakening the field on that side. Therefore, the electron path bends to the right (clockwise). If the permanent magnetic field strength is increased, the electron path will bend sharper. Likewise, if the velocity of the electron increases, the field around it increases and its path will bend more sharply.

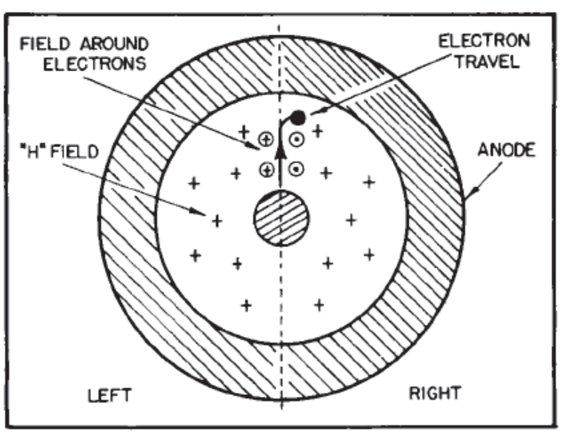


Figure 53-4 - Electron motion in a magnetic field.

A schematic diagram of a basic magnetron is shown in Figure 53-5.

The tube consists of a cylindrical plate with a cathode placed coaxially with it. The tuned circuit (not shown) in which oscillations take place are cavities physically located in the plate.

When no magnetic field exists, heating the cathode results in a uniform and direct movement in the field from the cathode to the plate, as illustrated in Figure 53-5B. However, as the magnetic field surrounding the tube is increased, a single electron is affected as shown in Figure 53-6. In Figure 53-6A, the magnetic field has been increased to a point where the electron proceeds to the plate in a curve rather than direct path.

In Figure 53-6B, the magnetic field has reached a value great enough to cause the electron just to miss the plate and return to the filament in a circular orbit. This value is the critical value of field strength. In Figure 53-6C, the value of the field strength has been increased to a point beyond the critical value, and the electron is made to travel to the cathode in a circular path of smaller diameter.

Figure 53-6D shows how the magnetron plate current varies under the influence of the varying magnetic field. In Figure 53-6A, the electron flow reaches the plate, so that there is a large amount of plate current flowing. However, when the critical field value is reached, as shown in Figure 53-6B, the electrons are deflected away from the plate; and the plate current drops abruptly to a very small value. When the field strength is made still larger, Figure 53-6C, the plate current drops to zero.

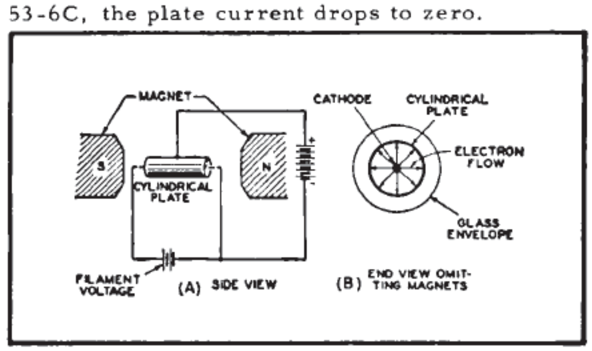


Figure 53-5 - Basic magnetron.

A1. The cavities.

A2. A magnetic field in the space between the cathode and plate of the magnetron serves as a grid to control the flow of plate current.

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When the magnetron is adjusted to the plate-current cut-off or critical value, and the electrons just fail to reach the plate in their circular motion, the magnetron can produce oscillations at an ultra-high frequency by virtue of the currents induced electrostatically by the moving electrons. This frequency is determined by the time it takes the electrons to travel from the cathode toward the plate and back again. A transfer of ultra-high frequency energy to a load is made possible by connecting an external circuit between the cathode and plate of the magnetron. Magnetron oscillators are divided into two classes, NEGATIVE RESISTANCE and ELECTRON RESONANCE magnetron oscillators.

A negative resistance magnetron oscillator operates by reason of a static negative resistance between its electrodes and has a frequency equal to the natural period of the tuned circuit connected to the tube.

An electron resonance magnetron oscillator operates by reason of the electron transit time characteristics of a vacuum tube, that is, the time it takes electrons to travel from cathode to plate. This oscillator is capable of generating very large peak power outputs at frequencies in the thousands of megacycles. Although its average power output over a period of time is low, it can put out very high power oscillations in short bursts of pulses; and is well adapted for pulse radars.

Q3. Will an electron travelling against the E lines in an electric field gain or lose energy ?

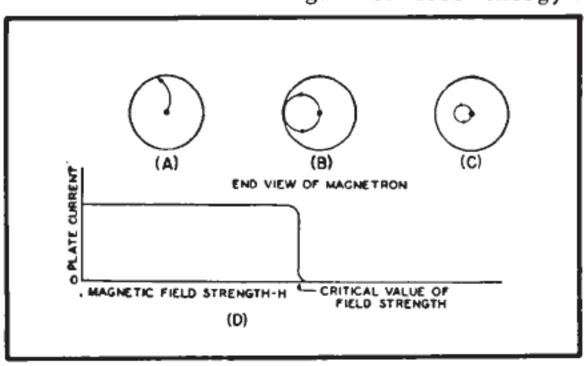


Figure 53-6 - Effect of magnetic field on single electron.

Q4. What causes an electron to travel in a curved path when it is moving perpendicular through an H field?

Q5. How do the electrons travel in a magnetron when the critical H field strength is applied?

## 53-3. Negative Resistance Magnetron

The split-anode negative resistance magnetron is a variation of the basic magnetron which operates at a higher frequency, and is capable of more output. Its general construction is similar to the basic magnetron, except that it has a split plate as shown in Figure 53-7. These half plates are operated at different po-

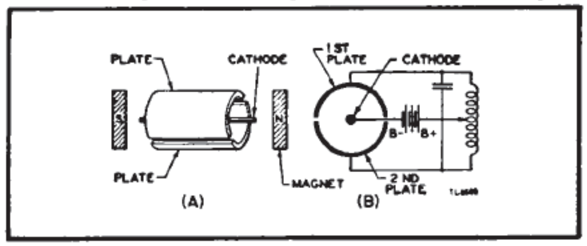


Figure 53-7 - Split-anode magnetron and its output.

tentials to provide an electron motion as shown in Figure 53-8. The electron leaving the cathode and progressing toward the high potential plate is deflected by the magnetic field at a certain radius of curvature. After passing the split between the two plates, the electron enters the electrostatic field set up by the lower potential plate.

Here the magnetic field has more effect on the electron, which is deflected at a smaller radius of curvature. The electron then continues to make a series of loops through the magnetic field and electric field until it finally

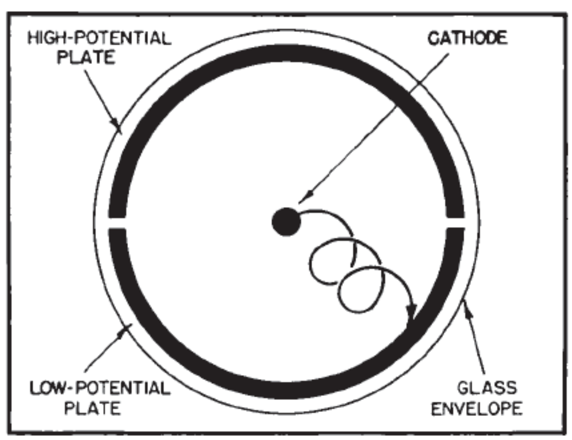


Figure 53-8 - Movement of electron in a split-anode magnetron.

falls on the low potential plate.

Oscillations can be started by applying the proper value of magnetic field to the tube. The value of field required is somewhat beyond the critical value which, for the split anode tube, is the field required to cause all electrons to miss the plate when its halves are operating at the same potential. However, the alternating voltages impressed on the plates as a result of the oscillation generated in the tank circuit will cause electron motion such as that shown in Figure 53-8, and current will flow. Since a very concentrated magnetic field is required for the negative resistance magnetron oscillator, the length of the tube plate is limited to a few centimeters for a magnet of reasonable dimensions. In addition a small diameter tube is required to make the magnetron operate efficiently at ultra-high frequencies. For this reason, a heavy-walled plate is used to increase the radiating properties of the tube. To obtain still greater dissipation, tubes with high outputs use an artificial cooling method such as forced air or water cooling.

The output of magnetrons is somewhat reduced by the bombardment of the filament by electrons which travel in loops 53-6B and C. This effect causes an increase of filament temperature under certain conditions of high magnetic field and high plate voltage; and sometimes results in unstable operation of the tube. The effects of filament bombardment can be compensated for by operating the filament at reduced voltage. In some cases, the plate voltage and field strength also are reduced to prevent destructive filament bombardment.

Q6. Why is the magnetron filament voltage often reduced after the magnetron is operating?

## 53-4. Electron Resonance Magnetron

In the electron resonance type of magnetron, the plate itself may be so constructed as to resonate and function as a tank circuit. Thus, there are no external tuned circuits; power is delivered directly from the tube to a transmission line as shown in Figure 53-9. The tube constants and operating conditions are such that

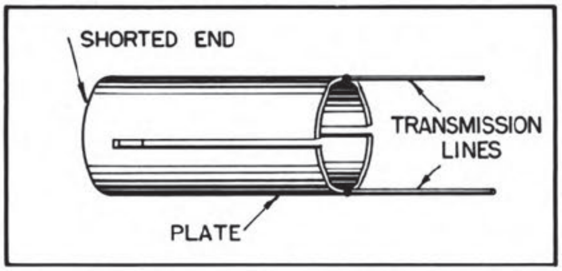


Figure 53-9 - Plate tank circuit of magnetron.

the electron paths are somewhat different from those in Figure 53-8. Instead of having closed spirals or loops, the path is a curve having a series of abrupt points as illustrated in Figure 53-10. Ordinarily, this type of magnetron also

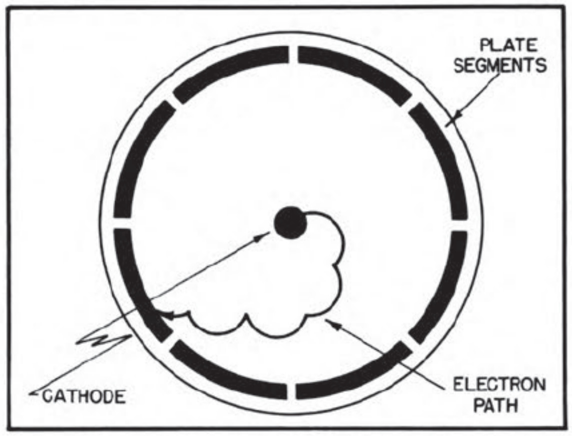


Figure 53-10 - Electron path in electron resonance magnetron.

has more than two segments in the plate. For example, Figure 53-10 illustrates an eight-segment plate.

This type of magnetron is the most widely used at present for ultra-high and super-high frequencies. Modern designs have a resonably high efficiency and relatively high output. However, one disadvantage of the electron resonance magnetron is that its average power is limited by the cathode emission. Furthermore, the peak power is limited by the maximum voltage which it can withstand without injury. Three common types of anode blocks used in electron-resonance magnetrons are shown in Figure 53-11.

The first type shown in Figure 53-11 has cylindrical cavities. This type is called a HOLE-AND-SLOT anode. The second type is called the VANE anode which has trapezoidal

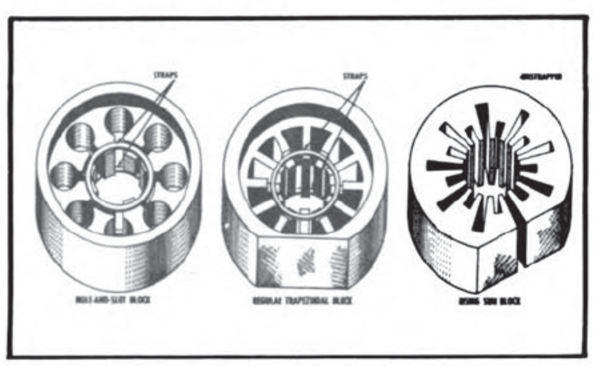

Figure 53-11 - Common types of anode blocks.

A3. The electron will gain energy.

A4. The interaction between the field about the electron and the H field.

A5. The electrons travel in circles just missing the anode and returning to the cathode.

A6. To reduce the effects of filament bombardment.

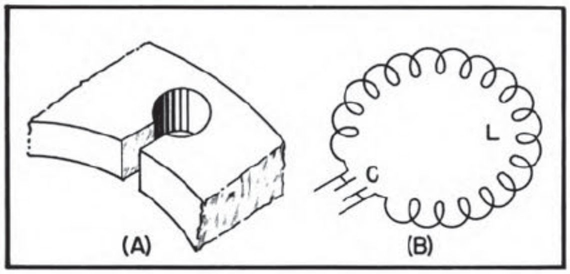


Figure 53-13 - Equivalent circuit of a hole and slot cavity.

cavities. These first two anode blocks operate in such a way that alternate segments must be connected, or strapped to insure that each segment is opposite in polarity to its neighboring segment on either side as shown in Figure 53-12. This also requires an even number of cavities.

The third type, illustrated in Figure 53-11, is called a RISING SUN block because of its appearance. The alternate large and small trapezoidal cavities in this block results in a stable frequency between the resonant frequencies of the large and the small cavities.

Figure 53-13A shows the physical appearance of the resonant cavities contained in the hole and slot anode which we will use when analyzing the operation of the electron resonance magnetron.

Notice that the cavity consists of a cylindrical hole in the copper anode, and a slot which connects the cavity to the interaction space.

The electrical equivalent circuit of the cavity and slot is shown in Figure 53-13B. The parallel sides of the slot form the plates of a capacitor, while the walls of the hole act as an inductor. The hole and slot thus form a high Q resonant LC circuit. As shown in Figure 53-11, the anode of a magnetron contains a number of these cavities.

An analysis of the anode in Figure 53-11 would reveal that the LC tank of each cavity are in series as shown in Figure 53-14. This is assuming the straps had been removed. However, an analysis of the anode block after alternate segments had been strapped in Figure 53-12 will reveal that the cavities are now connected in parallel. This is due to the strapping. The result is shown in Figure 53-15.

Q7. What is the equivalent circuit of a strapped anode block?

## 53-5. Operation of the Electron Resonant Magnetron

The electric field in the electron resonant oscillator is a product of an ac and a dc field. The dc field, previously described, extends radially between adjacent anode segments by the RF oscillations induced in the cavity tank circuits of the anode block.

Figure 53-16 shows the ac fields between adjacent segments at an instant of maximum mag-

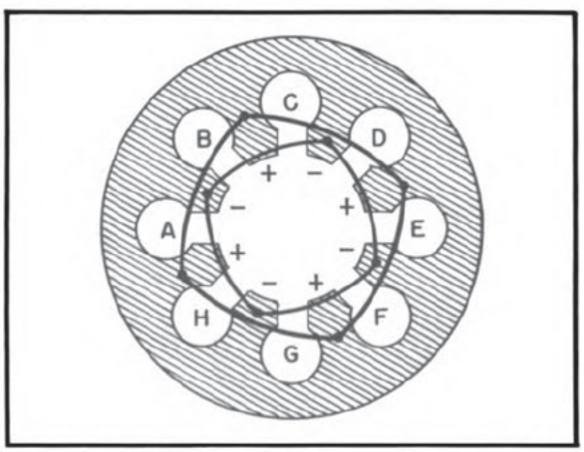


Figure 53-12 - Strapping alternate segments.

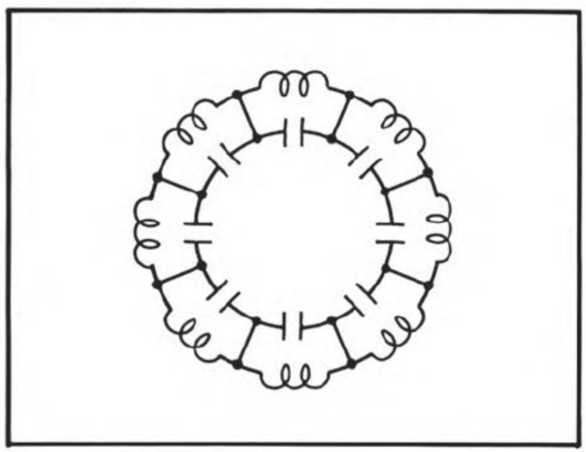

Figure 53-14 - Cavities connected in series.

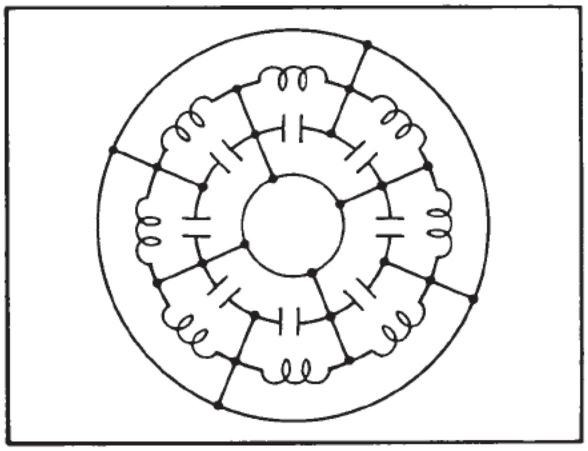

Figure 53-15 - Cavities in parallel due to strapping.

nitude of one alternation in the RF oscillations occurring in the cavities.

The strong dc field going from anode to cathode, due to a large negative dc voltage pulse applied to the cathode is not shown in Figure 53-16, but is assumed to be present. It is actually this strong dc field which causes electrons to accelerate toward the plate (anode) after they have been emitted from the cathode. Earlier it was pointed out that an electric (E) field went from a positive electrode to a negative electrode. Also, an electron moving against an E field would be accelerated by the field thus taking energy from the field. An electron would give up energy to a field and slow down if it were moving in the same direction as the field (positive to negative). Oscillations are sustained in a magnetron because electrons gain energy from the dc field, and give up this energy to the ac fields as they pass through these fields. These electrons are sometimes referred to as WORKING electrons. However, not all of the electrons give up energy to the ac fields. Some

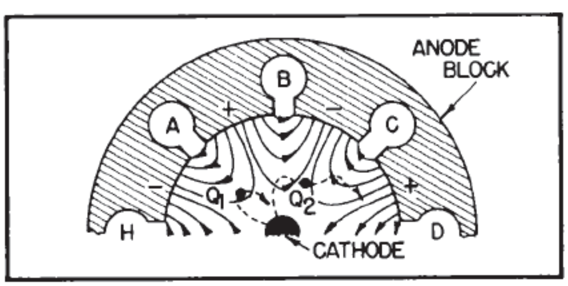


Figure 53-16 - Probable electron paths in an electron resonant magnetron oscillator—RF oscillations occurring.

electrons may actually take energy from the ac fields. This action is undesirable.

In Figure 53-16 consider electron $Q_1$, which is shown entering the field around the slot entrance to cavity A. The clockwise rotation of the electron path is due to the interaction of the magnetic field around the moving electron with the permanent magnetic field (H field) which is assumed to be going into the paper in Figure 53-16. The action of an electron moving in an H field was explained earlier and illustrated in Figure 53-4. Notice that electron $Q_1$ which has entered the ac field around cavity A is going against this ac field. Thus, it will take energy from the ac field and be accelerated. The electron will turn more sharply when its velocity increases as was explained earlier. Thus, electron $Q_1$ will turn back toward the cathode. When it strikes the cathode, it will give up the energy it received from the ac field in the form of heat. This will also force more electrons to leave the cathode and accelerate toward the anode. Electron $Q_2$ is, therefore, slowed down by the field and gives up some of its energy to the ac field. Since electron $Q_2$ looses velocity, the deflective force exerted by the H field is reduced and the electron path deviates to the left in the direction of the anode, rather than return to the cathode as did the former electron $Q_1$.

The cathode to anode potential and the magnetic field strength (E field to H field relationship) determines the time taken by the electron $Q_2$ to travel from a position in front of cavity B to a position in front of cavity C. Cavity C is equal to approximately one-half cycle of the RF oscillation of the cavities. When electron $Q_2$ reaches a position in front of cavity C, the ac field of cavity C will be reversed from that shown.

Therefore, electron $Q_2$ will give up energy to the ac field of cavity C and will slow down more. Electron $Q_2$ will actually give up energy to each cavity as it passes and will eventually reach the anode when its energy is expended. Thus electron $Q_2$ will have helped sustain oscillations because it has taken energy from the dc field and given it to the ac fields. Electron $Q_1$ which took energy from the ac field around cavity A did little harm because it immediately returned to the cathode.

Electrons such as $Q_2$ which give up energy to the ac field as they rotate clockwise from one ac field to the next are called working electrons, and stay in the interaction space for a considerable time before striking the anode.

The commulative action of many electrons being returned to the cathode and directed toward the anode forms a pattern resembling the spokes of a wheel. Since the motion of an individual electron in the field is in the clockwise

A7. A group of resonant circuits connected in parallel.

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direction as indicated in Figure 53-17. Electrons in the spokes of the wheel are the working electrons.

This overall space charge wheel rotates about the cathode at an angular velocity of two poles (anode segments) per cycle of the ac field, and of a phase that enables the concentration to continuously deliver energy to sustain the RF oscillations. Electrons emitted from the area of the cathode between the spokes are, as previously explained, quickly returned to the cathode.

In Figure 53-17 it is assumed that alternate segments between cavities are at the same potential at the same instant and that there is an ac field existing across each individual cavity. This type of mode of operation is called the pi mode, since adjacent segments of the anode have a phase difference of 180$^\circ$ or one pi radian. There are several other possible modes of oscillation.

A magnetron operated in the pi mode will have greater power output. Therefore, the pi mode is the most commonly used mode.

In order to insure that alternate segments have identical polarities, an even number of cavities, usually six or eight, are used and alternate segments are strapped as pointed out earlier. The frequency of the pi mode is separated from the frequency of the other modes by strapping. Operation in the pi mode is insured as follows:

For the pi mode, all parts of each strapping ring are at the same potential; but the two rings have alternately opposing potentials as shown in Figure 53-18. Stray capacitance between the rings adds capacitive loading to the resonant cavities, which lowers the frequency of the pi mode. For other modes, however, there is a phase difference between the successive segments connected to a given strapping ring which causes current to flow in the straps.

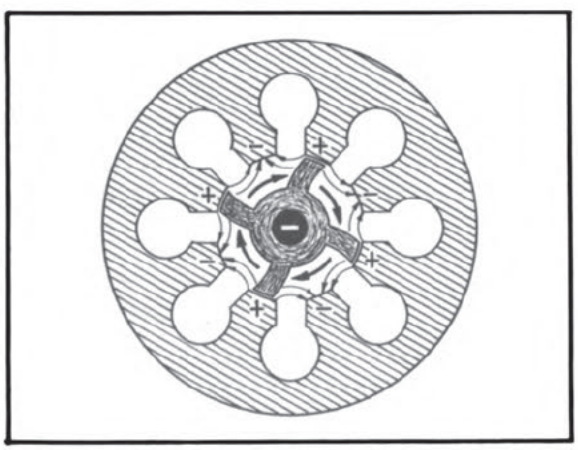

Figure 53-17 - Rotating space charge wheel in eight cavity magnetron.

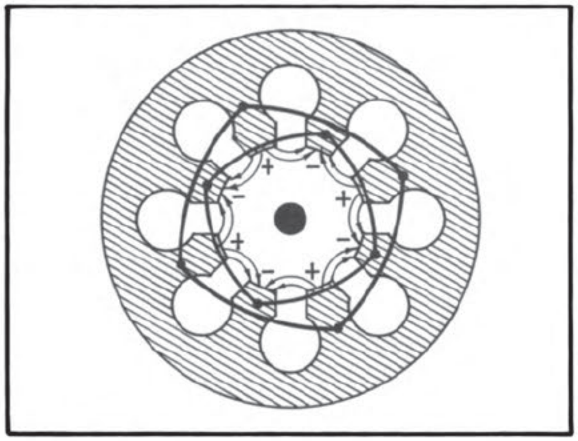

Figure 53-18 - Alternate segments connected by strapping rings.

The straps contain inductance and an inductive shunt is placed in parallel with the equivalent circuit thereby lowering the inductance and increasing the frequency at modes other than the pi mode.

The oscillations in the cavities are coupled out by means of a probe or loop or a slot extending into the cavity as shown in Figure 53-19.

Figure 53-20 shows a schematic diagram of a magnetron oscillator. Note that the plate or

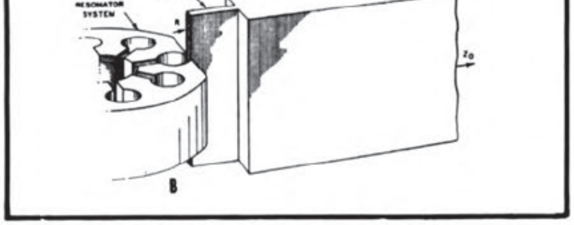


Figure 53-19 - Methods of coupling energy out of a magnetron.

outer shell is grounded for safety and that a high power negative dc pulse is applied to the cathode. The voltage developed across resistor $R_1$ when the magnetron conducts, causes a small current flow through the dc milliammeter, $M_1$. Capacitor $C_1$ filters the dc component from the pulse current flowing through $R_1$ so that a steady current flows through $M_1$.

The reading on $M_1$ indicates the average plate current drawn by the magnetron and is a valuable aid to the technician.

The current through the magnetron is very high, and it would not be practicable to put a meter in series with the conduction path of the magnetron. Therefore, $M_1$ is put in parallel with the small resistor, $R_1$, through which practically all of the magnetron plate current flows. There are many other methods of connecting a magnetron plate current meter into the circuit though only one method is shown in Figure 53-20.

Q8. How are electrons that take energy from the ac field eliminated?

Q9. On what three factors does oscillation in a magnetron depend?

Q10. How is energy coupled out of a magnetron?

## 53-6. Tuning and Stabilization

Magnetrons are inclined to stray from their proper frequency even under the most ideal conditions. Normally, this straying does not make much difference because the automatic frequency control in the radar receiver adjusts the frequency of the local oscillator to compensate for these variations.

Most military radar transmitters use TUNABLE magnetrons. These magnetrons have

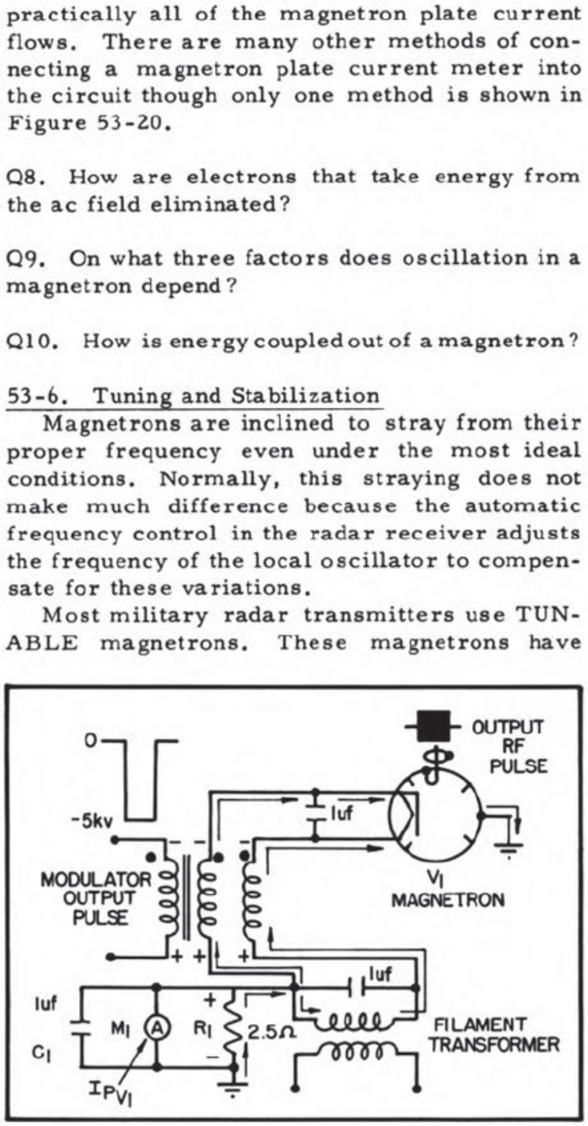


Figure 53-20 - Schematic diagram of a magnetron oscillator.

devices that permit the frequency to be tuned over a narrow range.

Until recently, magnetrons were operated on fixed frequencies. However, mechanical and electrical methods of tuning have proved successful over ranges up to 10 percent of the basic operating frequency of the magnetron.

Mechanical tuning can be defined as any method of tuning that does not involve the injection of additional electrons into the RF field. It generally takes the form of changing the capacitance or the inductance of one or more of the tuned cavities in the anode block. However, mechanical tuning can also be done by adding an external cavity or tuning stub in the output line.

VANE tuning is an example of one of these mechanical tuning methods. Small vanes are inserted into the end slots leading into the resonant cavities of the magnetron, as shown in Figure 53-21. These vanes, or plates, are inserted into or withdrawn from the end slots by means of a set of gears connected to a shaft extending into the anode block and attached to the plates. Because the surfaces of the end slots act as the plates of a capacitor, the insertion of the new conducting surfaces between the slots changes their capacity thereby changing the frequency.

PLUG tuning is achieved by the insertion of plugs into the resonant cavities as shown in Figure 53-22. This changes the inherent inductance of each cavity, and therefore, the resonant frequency is changed. A gear arrangement similar to that just described transmits the external motion into the magnetron interior.

In some tunable magnetrons, circular U-shaped metal bands are used for tuning. These bands are usually placed in special slots around the top of the anode block within the magnetron. They may be inserted into or withdrawn from the slots by a gear arrangement similar to that used for the plug or vane methods.

Electronic tuning includes all methods of tuning which involve the addition of electrons

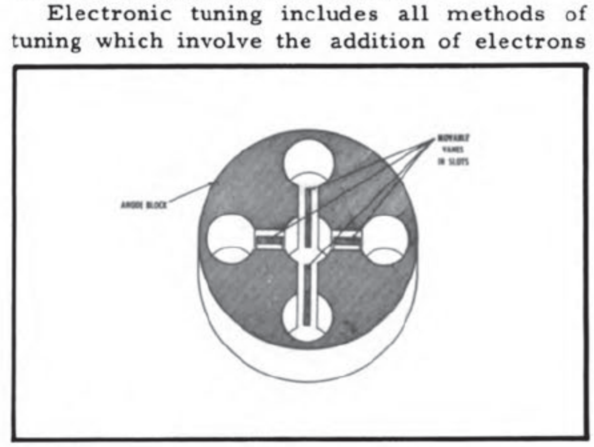

Figure 53-21 - Vane tuned magnetron.

A8. They are accelerated and returned to the cathode.

A9. Strength of the magnetic field, strength of the dc electric field, and the existance of resonant cavities or tuned circuits.

A10. RF energy is coupled out of a magnetron by a probe, loop or slot extending into one of the cavities.

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to the RF electron beam set up in the magnetron. This type of tuning is accomplished by adding an extra metal tube in the magnetron output. This tube may have a grid, or it may be a diode. Most tuning tubes are connected directly to the side of the magnetron.

Tuning is accomplished by passing the magnetron output through a beam of electrons in the tuning tube. As the intensity of the beam is changed, the magnetron output is frequency modulated.

Q11. Name two methods of tuning a resonant cavity.

STABILIZATION is considered with tuning because the two processes are related. Stabilization of the magnetron frequency is necessary in special cases when the frequency needs to be limited to a very small range or when the Q of the magnetron tanks is quite low.

Stabilization is essentially an energy-storage process. The magnetron is a poor device for storing RF energy. The stabilizer, on the other hand, presents a high Q at the frequency of the magnetron, and is therefore an efficient storage chamber.

Most stabilizers take the form of tuned cavities connected in the output line a few wavelengths from the magnetron. As the coupling between the low-Q stabilization cavity is critical at the radar frequency, some magnetrons are designed with a special built-in RF lead for the stabilizing cavity.

The stabilizing cavity itself is a tuned RF tank designed to present a low impedance to all frequencies except the magnetron frequency. Thus, when the magnetron output is coupled into the stabilizer, the stabilizer resonates at the fundamental frequency, and the several side-band frequencies are damped out. Energy from the resonating stabilization cavity is utilized to reinforce the magnetron output. The energy may, therefore, be considered as regenerative feedback from an equivalent circuit like the one shown in Figure 53-23. Since the stabilizer effectively reinforces the fundamental frequency and damps out the undesirable side-band frequencies, it can be considered as a simple form of automatic frequency control for the magnetron.

## 53-7. Checking for Proper Operation

The output circuits coupled to magnetrons are tuned to a specified frequency with bandpass characteristics equivalent to the operating frequency range of the magnetron. If the magnetron operates at a different frequency, it is not operating at maximum efficiency. The result may be a drop in magnetron plate current. Therefore, a frequency shift may show up as a drop in plate current that can be observed on the plate current meter. The simplest and surest check for the proper operating frequency of a magnetron, then, is to watch the plate current meter. The radar instruction manual gives an approximation, in milliamperes, of the best operating reading. In addition, daily readings should be taken and recorded. The average of these daily readings is then useful for comparison with the operating readings to determine whether the magnetron is operating normally. The readings should not be taken immediately after energizing the system because the magnetron requires a short period of time to warm up and become stable. Readings are also unstable for a short time after installing a new magnetron due to the seasoning time required for proper break-in.

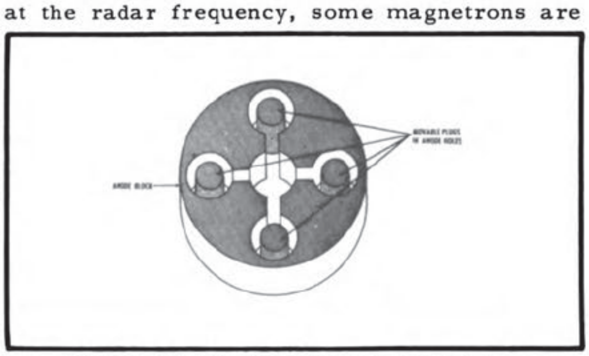

Figure 53-22 - Plug tuned magnetron.

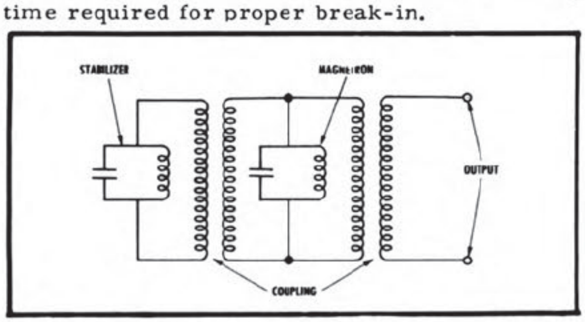


Figure 53-23 - Equivalent circuit of magnetron and stabilizer.

Q12. When would the plate current reading be taken on a magnetron?

If there is a mismatch in the output line, or waveguide, incorrect frequencies may also give maximum readings on the plate current meter, and the efficiency and range of the radar are lowered. Therefore, a device is needed to measure the magnetron output frequency directly. Such a device is the tunable ECHO BOX.

The echo box is a tunable resonant cavity connected to a crystal rectifier and a milliammeter. A portion of the transmitter pulse is coupled out of the waveguide and is fed into the cavity. The cavity is then tuned to resonance at the radar frequency, and this frequency can be read directly from the dial or from a calibration chart giving dial readings as a function of frequency.

Another use of the echo box is in making a SPECTRUM ANALYSIS of the radar pulse. A radar pulse consists not of a single frequency but of a band of frequencies centered in one fairly small range. A pulse spectrum is shown in Figure 53-24. The spectrum is obtained by tuning the echo box over the entire range of frequencies contained in the radar pulse and plotting the meter readings at intervals on a graph. It can be used in several ways. When a radar is first installed, a spectrum analysis is usually made to insure that excessive power is not dissipated in the side-band frequencies of the pulse. A poor spectrum, with large side bands, indicates a defective magnetron, a mismatch in the output line, or a poorly shaped modulator pulse.

ARCING in a magnetron is a warning sign that must be watched for. Arcing shows as bright flashes of light across the magnetron input leads. It can be detected by a characteristic singing sound, somewhat similar to the sound made by the tire of a semitrailer travelling at a high speed on a smooth pavement.

Most radars have a spark gap connected across the magnetron or the pulse transformer so that if arcing does occur, it will not occur in the magnetron. A small amount of arcing is to be expected, especially when the radar is first turned on or when a new magnetron has just been installed. However, the electronics technician must learn from experience when arcing is dangerous and when it is temporary and harmless.

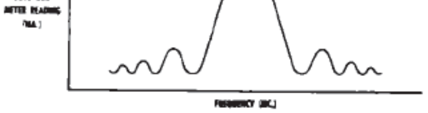

Figure 53-24 - Frequency spectrum of a radar pulse.

Arcing can be caused by several troubles. One common cause of arcing is allowing the input voltage to become too high. When high voltage arcing occurs, it can be stopped by lowering the input voltage until the plate current meter reading is normal. If the arcing is not stopped soon after beginning, the cathode may be destroyed.

Another cause of arcing is a physical obstruction or mismatch in the output line which sets up standing waves and causes RF energy to be reflected back to the magnetron.

When coupling is adjusted properly, the standing waves are removed and the arcing ceases. In a radar having a crystal protection, or SHUTTER relay in the waveguide, arcing sometimes means that the relay is locked shut or is malfunctioning, thus obstructing the passage of RF down the waveguide. When the relay is opened or replaced, the arcing should stop.

An aging magnetron can cause both arcing and low output, as indicated by the plate current meter.

If the cathode or one of the straps is burning out, or if one of the straps has shorted, arcing will occur. Therefore, if arcing occurs together with a high plate current reading and cannot be stopped by lowering the input voltage, the magnetron should be replaced. If either of the magnetron input leads are bent so that the filament is shorted to the surrounding fittings or structure, arcing is almost certain to occur. This cause of arcing is usually easy to detect because the plate current meter reading will be erratic.

Q13. What is the purpose of an echo box?

Q14. To what type of modulator is a magnetron most readily adapted?

Q15. What are the most common causes of arcing in a magnetron?

## 53-8. Troubleshooting

Radar troubleshooting involves not only the replacement of a component which is obviously malfunctioning, but it also involves the ability to recognize a slight decrease in optimum performance; and correct it before serious trouble developes. To do this, periodic routine checks of meter indications must be made and recorded.

For example, assume that the magnetron plate current readings have been decreasing over the past three readings. These readings

A11. Two types of tuning a resonant cavity are vane and plug tuning. Both of these are mechanical.

A12. Magnetron plate current readings should not be taken until the magnetron has been allowed to reach its operating temperatures.

A13. An echo box is used to accurately measure the magnetron output frequency. It can also be used to determine the power dissipation in the side-band frequencies.

A14. The magnetron oscillator is most readily adapted to pulse modulation.

A15. Arcing can be caused by too high an input voltage pulse, by a mismatch in the transmission line, or shorted line.

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indicated that the trouble must lie somewhere in the transmitter circuit. By knowing the function of each block in the transmitter unit, the trouble can be isolated to one of three causes: the magnetron is defective, the input pulse is poorly shaped, or standing waves have been set up because of a mismatch in the waveguide or output line. Rule out the last of these three causes because the drop in the readings has been steady. A mismatch gives a drop or rise when it first occurs. Thus, the trouble can be narrowed down to two probable causes. As the input pulse to the magnetron is of high power, it is difficult to measure and determine if it is of the proper voltage. If the output pulse from the modulator is faulty, the trouble is most likely the result of either a weak tube in the modulator power supply, a defective element in the pulse line, or a defective magnetron. Replacing the magnetron would take more time than running a test on the modulator. Therefore, the modulator would be examined before checking the magnetron.

The best practical way to check for a defective magnetron, when necessary, is to replace it. Shut down the equipment and ground the storage capacitor in the modulator and power supply. Follow the instructions given in the radar instruction manual to replace the magnetron. Handle the old one with care, it may still be operative. Pack it carefully in a box and store it in a dry, cool place. When installing the new magnetron, make sure the RF coupling is thoroughly clean and secure. Check the spaces on both sides of the magnetron between the anode block and each pole piece of the permanent magnet to insure that the magnetron is properly centered.

Remember to REMOVE YOUR WATCH when working near a magnet. Do not bring metal tools near the magnet, since even a slight blow from a metal tool can cause a serious loss in the field strength of the magnet. Non-magnetic tools should be used when possible. Always keep in mind that the magnetron is expensive and can be damaged easily.

After the new magnetron has been installed, double check to insure that all leads are tight with proper tolerances and that the magnetron is correctly centered. Turn on the radar main power and apply heater voltage. Allow the tube to warm-up for about 5 minutes before applying the high voltage. Most radar systems are equipped with a 5 minute time delay circuit for this purpose. After the warm-up, increase the input voltage slowly until the normal operating value is reached or until excessive arcing occurs. If arcing occurs, reduce the voltage slightly until the arcing stops and allow the magnetron to operate at the reduced voltage for about 5 minutes. Then build up the voltage until it reaches normal operating value or until arcing again occurs. This process should be repeated until the new magnetron is operating steadily at normal operating power. For the first 10 or 15 minutes slight arcing may be expected and variations in the output current may appear, because gas that has accumulated in the magnetron is burning out. This process is called SEASONING.

If the magnetron overload relay trips out when the high voltage is applied, reset the relay and repeat the process at least five times before a decision to replace the magnetron is made.

Q16. If a magnetron is suspected of being defective, what is the best method of checking it?

Q17. Why should metal tools be kept from contact with the magnetron magnet assembly?

Q18. What method is used to prevent high voltage from being applied immediately to the magnetron when main power is applied?

## DUPLEXERS

### 53-9. Principles of Duplexers

Separate antennas and transmission lines are sometimes used for transmitting and receiving RF signals. A system of this type is naturally more expensive and space consuming than a system that employs one antenna for both transmit and receive. For this reason, a device called a DUPLEXER has been developed which permits the use of a common transmission line and a

single antenna for transmitting and receiving. An important point to consider when using an antenna for both transmitting and receiving is keeping the transmitted signal out of the receiver. The high transmitter power output could damage the receiver to a point where it could not receive at all or cause a blocking action which would prevent the reception of the returned echo. Also, the low received signal power would be critically weakened, or lost entirely, if it were allowed to be absorbed in the transmitter output coupling.

A duplexer system consists of a TR (TRANSMIT-RECEIVE) and an ATR (ANTI-TRANSMIT-RECEIVE) circuits. The purpose of a TR circuit is to prevent the transmitter power from entering the receiver. The purpose of the ATR circuit is to prevent the received signal from entering the transmitter, so that all of the received signal is sent to the receiver. The transmission line in which the TR and ATR device are housed may be either a coaxial line or waveguide; all operate on the same principle.

Figure 53-25 shows two duplexer systems used to connect a radar transmitter and receiver to a common antenna. For simplicity, a two-wire

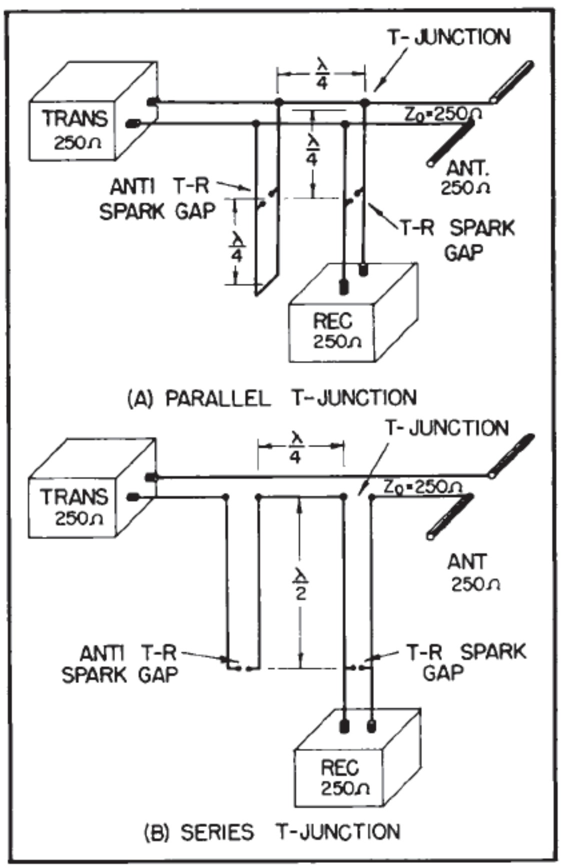


Figure 53-25 - Duplexer systems.

line will be used when analyzing the duplexer. The junction of the transmitter, receiver and antenna coupling transmission lines is called the T-JUNCTION. This T-junction may use a parallel connection as in Figure 53-25A or a series connection as in Figure 53-25B. The duplexer will be analyzed using the parallel connected T-junction first. The TR and ATR devices are spark gaps as shown in Figure 53-25A and B. The TR circuit consists of the TR device and the transmission line in which it is housed, likewise, the ATR circuit consists of the ATR device and the transmission line in which it is housed.

In order to analyze the over-all action of the TR-ATR circuits, recall that a quarter wavelength section of transmission line (or multiple thereof) inverts the impedance at the receiving end and a half wavelength (or even multiples of a quarter wavelength) repeats the impedance at the receiving end. The TR spark gap in Figure 53-25A is located in the receiver coupling one-fourth wavelength from the T-junction. One (or any odd multiple) quarter wavelengths from the T-junction, in the direction of the transmitter, a one-half wavelength, closed-end, section of transmission line, called a STUB, is attached to the main transmission line. An ATR spark gap is located in this line one-quarter wavelength from the main transmission line, and one-quarter wavelength from the closed end of the stub.

The spark gap makes a reasonably good switch because it is an open circuit until sufficient voltage is applied to cause the gap to arc over. The arc is formed by causing the gas or vapor between the electrodes to ionize. Once started, the running voltage across the gap is very low, and the resistance of the gap approaches a short circuit. The ionized gap voltage is independent of the applied power, so that the resistance varies with applied power. Air at atmospheric pressure requires about 30,000 volts per inch of gap to start the arc, while the running voltage is about 50 volts. The break-down and running voltages for inclosed gaps depend on the pressure and the gas or vapor used.

For purposes of illustration, it will be assumed that the characteristic impedance of the transmission line, the feed point resistance of the antenna, the input impedance of the receiver, and the output impedance of the transmitter when generating RF power, are all 250 ohms. The transmitter output impedance rises between pulses to 5,000 ohms. The resistance of the conducting gap is 50 ohms.

During the transmitting pulse, an arc appears across both spark gaps and causes the TR and ATR circuits to act as a shorted transmission line. However, an open circuit is reflected one-

A16. The best method of checking a magnetron is to replace it.

A17. Metal tools and objects should be kept from contact with the magnet to prevent the weakening of its field strength.

A18. In most systems, a delay circuit is incorporated for this purpose.

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quarter wavelength away, where the TR and ATR circuits are connected to the main transmission line. Figure 53-26 shows the circuit during transmission.

None of the transmitted energy can pass through these reflected opens into the ATR stub or into the receiver. Therefore, all of the transmitted pulse is sent straight to the antenna. Remember that the antenna impedance, the line impedance, and the transmitter impedance, when transmitting, are all matched.

During reception, the amplitude of the received echo is not sufficient to cause an arc across either spark gap. Under this condition, the ATR circuit now acts as a half-wave transmission line terminated in a short. This is reflected as an open circuit at the T-junction, three-quarter wavelengths away.

The received echo sees an open circuit in the direction of the transmitter as shown in Figure 53-27. However, the receiver input impedance is matched to the transmission line impedance so that the entire received signal will go to the receiver with a minimum amount of loss. The duplexer using a series connection at the T-junction, is somewhat different but the result is the same. Note, in Figure 53-25B,

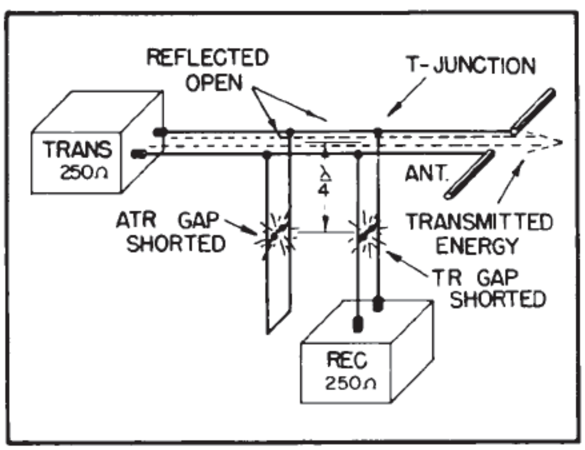


Figure 53-26 - Parallel connected duplexer during transmission.

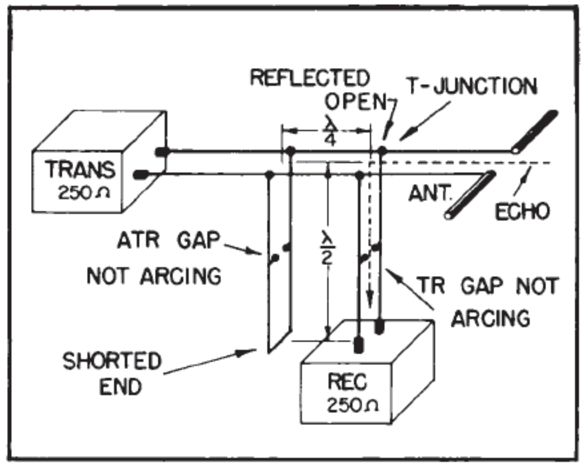


Figure 53-27 - Parallel connected duplexer system during reception.

that the ATR gap is one-half wavelength from the main transmission line.

During transmission, the ATR and TR gaps fire in the series connected duplexer system. However, this causes a short circuit to be reflected at the series connection to the main transmission line one-half wavelength away as shown in Figure 53-28.

The transmitted pulse sees a smooth path in the direction of the antenna, and does not go into the ATR stub or the receiver.

During reception, neither spark gap is fired. The ATR acts as a half-wave stub terminated in an open. This is reflected as a short circuit at the T-junction three-quarters of a wavelength away as shown in Figure 53-29. Consequently, the received signal sees a smooth path to the

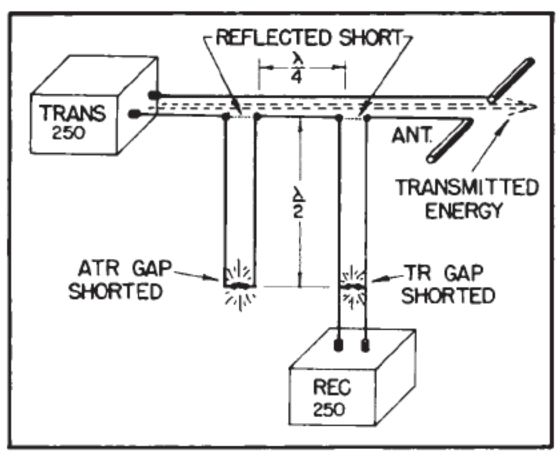


Figure 53-28 - Series connected duplexer system during transmission.

receiver, and none of the received signal is lost in the transmitting circuit.

Q19. What is the function of a duplexer system?

Q20. In a duplexer system using a parallel connected T-junction why are the ATR and TR devices an odd multiple of a quarter wavelength from the transmitting antenna?

Q21. What is the purpose of the TR circuit?

Q22. What is the purpose of the ATR circuit?

## 53-10. TR and ATR Tubes

It is necessary for the spark gaps to fire or ionize very fast when the transmitter pulse comes down the line. The spark gaps must also deionize immediately after the transmitter pulse has been transmitted so that received echos from very close targets can be received.

The spark gap used in a given TR system may vary from a simple one formed by two electrodes placed across the transmission line to one inclosed in an evacuated glass envelope with special features to improve operation. The requirements of the spark gap are that its resistance shall be very high until the arc is formed, and then be very low during conduction through the arc. At the end of the transmitted pulse, the arc should be extinguished as rapidly as possible to remove the loss caused by the arc, and to permit signals from nearby targets to reach the receiver.

The simple gap formed in air has a resistance during conduction of from 30 to 50 ohms. This is usually too high for use with any but an open-wire transmission line. The time required for the air surrounding the gap to be completely deionized after the pulse voltage has been removed is about 10 microseconds. During this time, the gap acts as an increasing resistance across the transmission line to which it is connected. However, in a TR system using an air gap, the received signals reaching the receiver through the gap have half their proper magnitude after 3 microseconds. This is known as RECOVERY TIME.

The value of voltage required to break down a gap, and the running voltage during the arc, can be lowered by reducing the pressure of the gas surrounding the electrodes. TR tubes are therefore used in which the spark gap is inclosed in a glass envelope and the tube is partially evacuated as shown in Figure 53-30. The arc is formed by conduction through an ionized gas or vapor so that the tube cannot be entirely evacuated; thus, there is an optimum pressure which will give the best TR operation. The recovery time or deionization of the gap can be reduced by introducing water vapor into the tube rather than air. A TR tube containing water vapor at a pressure of 1 millimeter of mercury will recover in 0.5 microseconds.

TR tubes for use at microwave frequencies are built to fit into and be a part of a resonant cavity or transformer. The high Q of the cavity and the vapor in the evacuated portion of the tube reduce the power needed to maintain the gap and the power of the transmitted pulse which reaches the receiver. The speed with which this action takes place can be increased by placing a third electrode within one of the main electrodes of the gap. This electrode is known as a KEEP ALIVE, and has a potential of about -1,000 volts with respect to the main gap. A glow discharge is maintained by the keep alive and one electrode of the main gap when the transmitted pulse is applied. The negative voltage of the keep alive also prevents stray ions from reaching the main gap to produce noises in the receiver. The negative keep alive potential will attract stray ions because the gases in the TR tubes are mainly composed of positive ions.

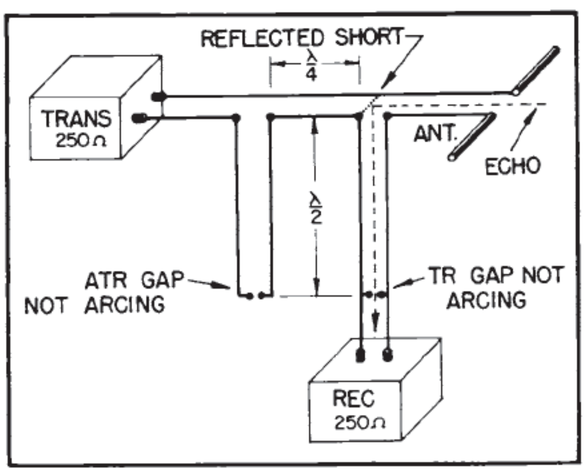


Figure 53-29 - Series connected duplexer system during reception.

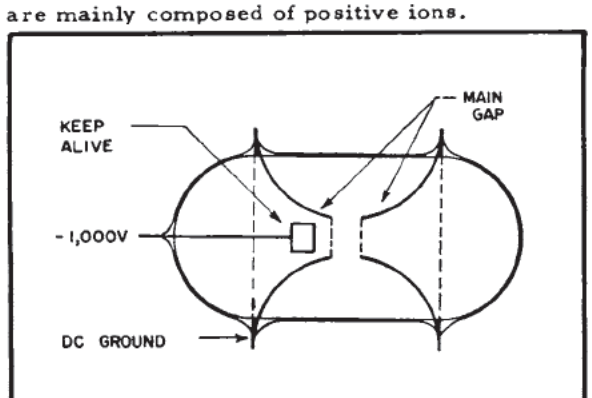


Figure 53-30 - TR tube with keep alive electrode.

A19. It permits the use of a common transmission line and antenna for transmitting and receiving.

A20. The TR and ATR devices are located at odd multiples of a quarter wavelength so that during transmission and reception proper impedances are reflected.

A21. The purpose of the TR circuit is to prevent the transmitted pulse from entering the receiver.

A22. The purpose of the ATR circuit is to prevent the received signal from going into the transmitter.

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The life of the TR tube is controlled by two main factors. The first and most common cause of failure is due to a gradual buildup of metal particles knocked loose from the electrodes of the gap and spattered on the inside of the glass envelope. These particles act as small, conducting areas which lower the Q of the resonant cavity and waste power. If the tube is continued in use for any length of time, the particles will begin to form a detuning wall within the cavity which will eventually prevent the TR tube from functioning properly. The second cause of failure is due to an absorption of the gas within the tube by the metal electrodes. The result is to reduce the pressure gradually within the tube to the point where it becomes very difficult to break down the gap, and extremely strong signals are fed to the receiver. Because both causes of failure are only gradually noticed, the TR tube must be checked carefully and periodically for efficient operation.

In numerous applications, TR tubes are mounted in cylindrical cavities, with the metal electrodes connected to and forming part of the end walls as shown in Figure 53-31. The cavity is excited in a mode which produces a strong electric field across the gap, and therefore causes an arc with the minimum applied signal. The method of exitation used in the cavity shown is to terminate a coaxial feed line within the cavity by forming a coupling loop with the inner conductor. The coupling loop is a low impedance across the coaxial line so that current through the loop is large, and a strong magnetic field is set up. The loop is placed so as to reinforce the magnetic field within the cavity. The amount of coupling is controlled by rotating the loop on an axis formed by the inner conductor. Signals for the receiver are removed from the cavity by a similar loop placed on the opposite side of the gap from the input loop.

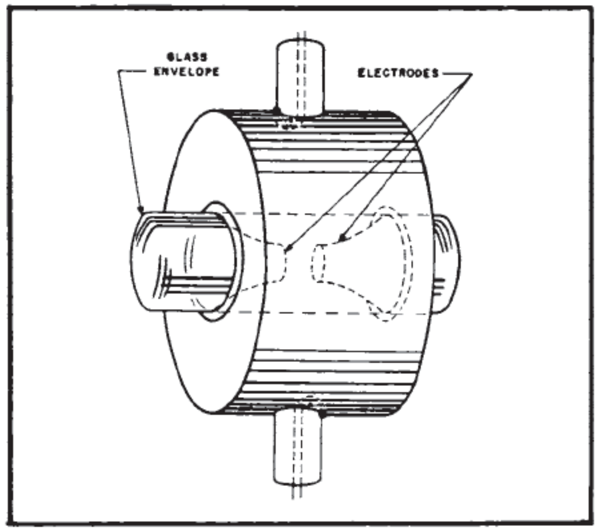


Figure 53-31 - Cavity TR box.

Another method of feeding the cavity from a coaxial line is to use slots which couple the field of the line to that of the cavity as shown in Figure 53-32. The output of the magnetron is matched into a coaxial line which feeds the transmitted pulse to the antenna. Near the magnetron, the outer conductor of the coaxial line is made into a sliding section having a side aperture or slot in the center. The slot opens into the resonant cavity of the TR box on the periphery. Another slot on the opposite side of the cavity is cut through into the receiver coaxial line, which is shorted at the edge of the slot.

During the transmitted pulse, energy is coupled into the cavity and produces a large voltage across the gap. The gap breaks down, forming an arc which short-circuits the center of the cavity. The field built up within the cavity is very weak because of the detuning of the cavity during the arc, and has the effect of a short circuit across the transmitter feed line slot. Because the field is weak, very little

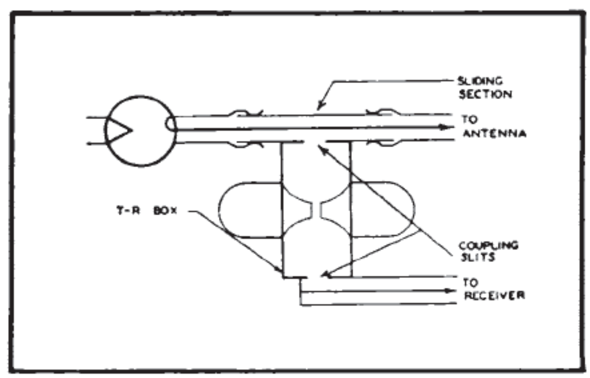


Figure 53-32 - Slot-coupled cavity.

energy is coupled into the receiver feed line.

At the end of the transmitted pulse, the magnetron impedance changes, so that the received signal sees a mismatch at the magnetron. The position of the slot into the cavity is adjusted by means of the sliding joint to be placed at a current maximum of the coaxial line as shown in Figure 53-33. The large current produces a strong magnetic field which leaks through the slot into the cavity to reinforce the field of the cavity. The received signal is not strong enough to break down the gap, and the field within the cavity is not affected by the presence of the gap. The slot into the receiver feed line permits some of the field of the cavity to link the inner conductor of the line at a point which is a current maximum. By selecting the proper size and shape of the slots, the received signal is passed, with very little loss, to the receiver. The position of the antenna feed line slot can be adjusted to absorb all of the received line energy from the antenna feed line, and therefore to prevent standing waves between the TR switch and the antenna.

Resonant cavity TR switches are applied to waveguides both directly or indirectly to obtain switching action. The indirect method uses a coaxial line into the waveguide which feeds the antenna. When large losses might be incurred by the use of coaxial line, the resonant cavity can be coupled directly to the waveguide. Figure 53-34 shows a direct method of cavity TR switching in a waveguide feed system. The waveguide terminates in the antenna on one end and a shorting plate at the other.

The transmitted pulse travels up the guide, spilling into the cavity through a slot. The cavity builds up a strong electric field across the gap, breaking it down, and detuning the cavity. The impedance seen at the slot by the guide is decreased to an approximate short circuit which effectively seals the opening and passes the pulse energy to the antenna.

The signals received during the resetting

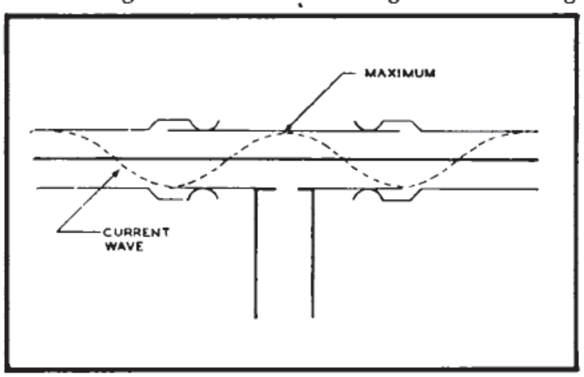


Figure 53-33 - Current wave reflected by magnetron.

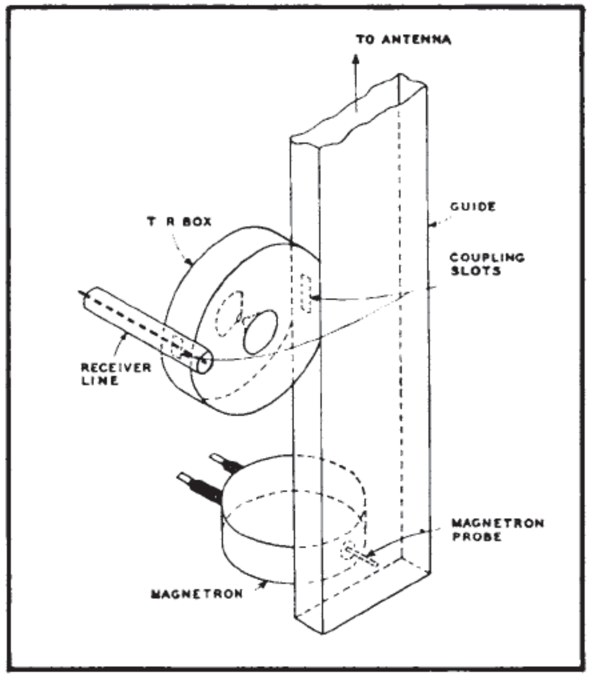


Figure 53-34 - Waveguide using cavity TR box.

time travel down the guide to the magnetron and the shorting end plate, where they are reflected. The slot into the cavity is placed so that it is located at a maximum of the standing wave magnetic field produced by reflections. The maximum field therefore links the magnetic field of the cavity. Since the received signals are not strong enough to cause an arc, the cavity field is undisturbed by the gap, and transfers into the receiver coaxial line to give maximum energy transfer.

The cavity TR switch can also be applied to branch lines of the waveguide as shown in Figure 53-35. The magnetron is coupled to the guide by a voltage probe to produce proper excitation. In order to insure maximum use of the received signals an ATR switch has been included. The transmitted pulse travels from the magnetron to the ATR branch where part of the energy turns into the gap. A slot is placed across the waveguide a half-wavelength from the main guide, and passes the RF energy through into the cavity. The cavity builds up the electric field, breaks down the gap, and as a result produces a short circuit across the coupling slot. The short circuit of the slot is reflected back to the main guide a half-wave away to close the entrance to the ATR branch.

Most of the energy is therefore directed down the guide to the antenna. On reaching the receiver branch, the same effect is produced by the TR switch a half-wavelength from the main

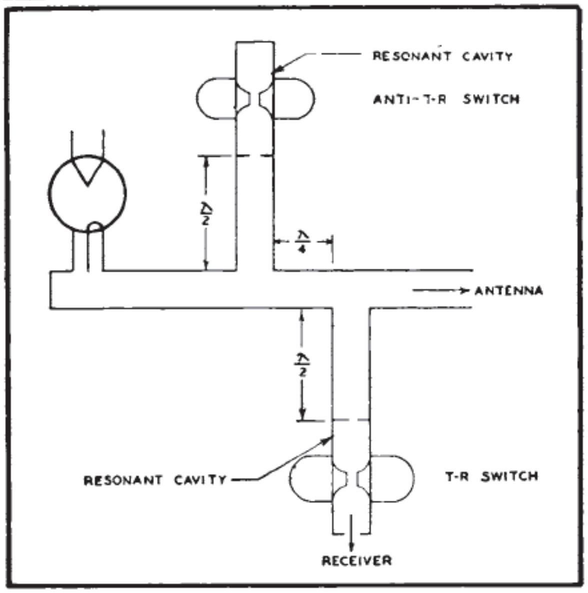


Figure 53-35 - Cavity TR box applied to branch lines of waveguide.

guide. Since both openings are effectively closed by the gaps, maximum energy is transferred between the magnetron and the antenna.

During the resting time, the ATR spark gap is not broken down by the received signals, so that the input to the cavity is practically an open circuit. This is reflected to the main guide as an open circuit. The received signals are in effect turned back by the apparent open circuit at the mouth of the ATR branch setting up reflections which a quarter-wave away at the TR branch produce a short or closing of the main guide. The signals are directed into the TR branch where they pass through the resonant cavity of the receiver.

Instead of using resonant cavities and TR tubes, the branch waveguide can use RESONANT SLOTS which also act as spark gaps. An example is illustrated in Figure 53-36. The resonant slot is a partition across the guide with an aperture whose dimensions make it look like a parallel resonant circuit at the carrier frequency. The dimension in the direction of the electric field is made small so that the transmitted pulse will cause an arc. The arc closes the conducting surface of the slot, providing a short circuit which is reflected by the half wave line to the main guide.

Q23. Why is it necessary for the TR tube to deionize instantly?

Q24. What is the purpose of the keep-alive voltage applied to the TR tube?

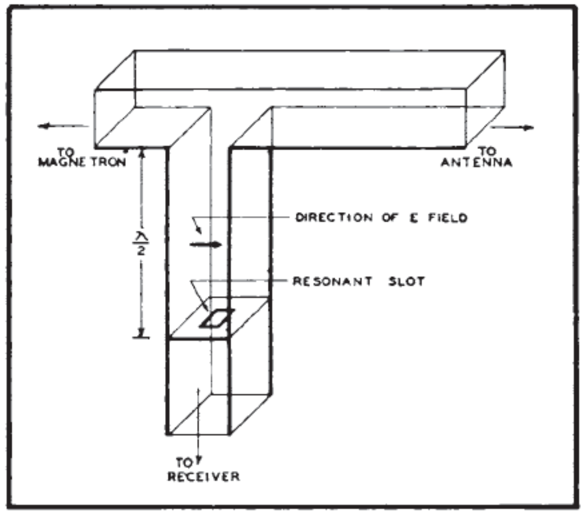


Figure 53-36 - Slot type TR switch.

Q25. Why is the polarity of the keep-alive voltage always negative?

Q26. What symptoms would be present if the TR tube failed to ionize?

## 53-11. TR and ATR Switch Transformer

Both the TR and the ATR switches require some of the transmitted pulse power to operate them. This is undesirable, since part of their function is to increase efficiency. The amount of power required can be somewhat reduced, and the switching action improved, by using transformers to step up the voltage applied across the gap. Suppose that the signal is applied to the primary of the step-up transformer, and the spark gap is placed across the secondary as shown in Figure 53-37. The secondary is tuned to obtain a very high impedance when the gap is not conducting. This impedance is reflected to the primary as an open circuit.

The voltage of the signal applied to the primary is stepped up in the secondary, which causes the spark gap to break down sooner than if the original signal were applied directly to the gap. The resistance of the conducting gap is placed across the tuned secondary, and is stepped down into the primary as a much lower resistance. Assuming a gap resistance of 50 ohms, and a step-down of 10:1, the primary will have an impedance of 5 ohms. The 5 ohms is reflected by the quarter-wave line as:

$$Z = \frac{250^2}{5} = 12,500 \text{ ohms}$$

The power taken from the gap will therefore be reduced by 10:1, as compared to placing the

gap directly across the line.

Transformers of the ordinary RF type will not function well at the carrier frequencies used in radar, while the resonant line can be used as an excellent auto-transformer to produce the same results. Figure 53-38A shows an open wire line which uses a quarter-wave stub to step up the voltage applied to the ATR spark gap and to step down its conducting resistance across the ATR line. The stub can be considered as two sections of transmission line, as shown in Figure 53-38B, one of which is terminated in a short circuit, and the other is an open circuit. The shorted line is less than a quarter-wavelength and acts as a capacitance to an inductance. The open line is also less than a quarterwave and acts capacitive.

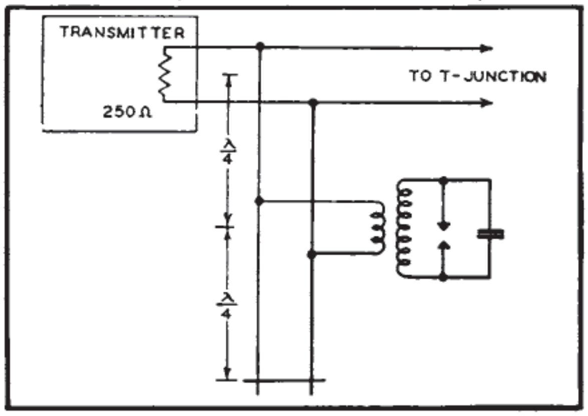


Figure 53-37 - ATR switch with transformer.

The two are in parallel, and since their total length is a quarter-wave, their reactances are equal, thus forming a parallel resonant circuit of very high impedance as shown in Figure 53-38C. During the resting time, with the gap extinguished, the stub high impedance has very little effect in bridging the ATR line.

When energy is applied to the stub, a standing wave of voltage that is maximum across the gap

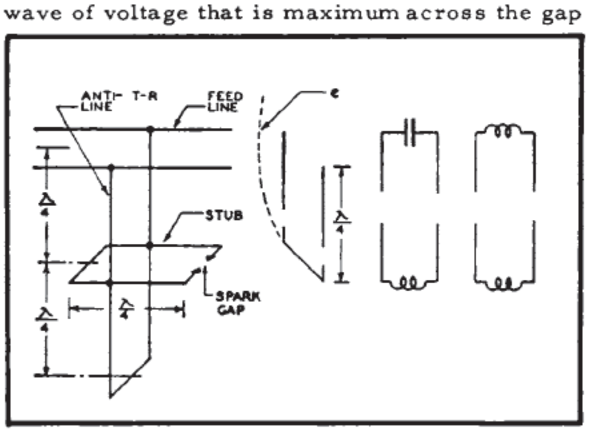


Figure 53-38 - Resonant line transformer ATR.

is set up along its length. The received signals are normally not large enough to break down the gap. The transmitted pulse is large enough, however, and places a low resistance across the open end of the stub causing the gap to conduct. The stub now consists of two lines in parallel across the ATR line, each of which is inductive as shown in Figure 53-38D. The result is to place a very low inductance across the ATR line a quarter wavelength from the feed line, reflecting the low impedance as a very high impedance to the feed line to limit the energy necessary to operate the gap.

Q27. What is the function of the resonant line transformer?

### 53-12. Handling and Disposal

Many TR tubes contain radioactive material and should be handled with care to avoid breakage. The radioactivity level of an unbroken tube is very minute and presents no danger to personnel during normal handling. All TR tubes, damaged or not, should be disposed of in accordance with BUSHIPS instruction 5100.5 and BUSHIPS manual, chapter 67, article 314.

## ANTENNAS

### 53-13. Principles of Radar Antennas

Radar antennas fall into two general classes—OMNI-DIRECTIONAL and DIRECTIONAL. Omni-directional antennas radiate RF energy in all directions simultaneously, as their name implies. They are seldom used with modern radars, but are commonly used in radio equipment, in IFF (Identification Friend or Foe) equipment, and in counter measures receivers for the detection of enemy radar signals. Directional antennas, on the other hand, radiate RF energy in LOBES or BEAMS, that extend outward from the antenna in one direction or in only a few directions for a given antenna position. Sometimes, the radiation pattern contains small minor lobes, but these lobes are weak and normally have little effect on the main radiation pattern. The main lobe may range in angular width from 1° or 2° in some radars to several degrees in other radars, depending on the accuracy demanded. The energy of the main lobe is restricted either in the horizontal or in the vertical plane, or in both, to a few degrees. The minor lobes are made as small as possible in order to form a maximum amount of energy in to the main lobe.

Directional antennas have two important characteristics. One is DIRECTIVITY. The directivity of an antenna refers to the degree

A23. The TR tube must deionize instantly so that an echo can be received from the close targets.

A24. The purpose of the keep-alive voltage is to aid in the rapid ionization of the TR tube.

A25. The negative polarity of the keep-alive voltage prevents stray ions within the TR tube from reaching the spark gap and producing noise in the receiver.

A26. The receiver would have a long recovery time.

A27. A resonant line transformer is frequently used to step up the voltage applied to the spark gap in the TR or ATR tube. This reduced the amount of energy taken from the transmitter pulse and also improves switching action.

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of sharpness of its beam. If the beam is narrow in either the horizontal or vertical plane, the antenna is said to have high directivity in that plane. Conversely, if the beam is broad in either plane, the directivity of the antenna in that plane is low. Thus, if an antenna has a narrow horizontal beam and a wide vertical beam, the horizontal directivity is high and the vertical directivity is low.

When the directivity of an antenna is increased, that is, when the beam is narrowed, less power is required to cover the same range because the power is not scattered over so wide an area. Thus, the second characteristic of an antenna is brought to light. This characteristic is called POWER GAIN, and is directly related to directivity.

Power gain is the ratio of power at some point in the radiation field of an antenna over the power at the same point of a single dipole antenna located in the same position and fed in the same way as the antenna being measured. An antenna with high directivity has a high power gain, and vice versa. The power gain of a single dipole with no reflector is one. An array of several dipoles in the same position as the single dipole and fed with the same line would have a power gain of more than one, the exact figure depending on the directivity of the array.

The BEARING RESOLUTION of a radar system depends on the width of the antenna beam. Resolution is the ability of the radar equipment to distinguish small targets or the shape of a large target. To distinguish any of the details of a large target, the beam must be much smaller in cross section than the target and the pulse length must be very short.

The measurement of the bearing of a target as "seen" by the radar is usually given as an angular position. The angle may be measured either from true north (true bearing), or with respect to the heading of a vessel or aircraft containing the radar set (relative bearing). The angle at which the echo signal returns is measured by utilizing the directional characteristics the radar antenna system. Radar antennas are constructed of radiating elements, reflectors, and directors to produce a single narrow beam of energy in one direction. The pattern produced in this manner permits the beaming of maximum energy in a desired direction. The transmitting pattern of an antenna system is also its receiving pattern. An antenna can therefore be used to transmit energy, to receive reflected energy, or to do both.

The simplest form of antenna for measuring azimuth or bearing is one that produces a single-lobe pattern. The system is mounted so that it can be rotated. Energy is directed across the region to be searched, by moving the beam back and forth in azimuth until a return signal is picked up. The position of the antenna is then adjusted to give maximum return signal.

Figure 53-39 shows the receiving pattern for a typical radar antenna. In this figure, relative signal strength is plotted against the angular position of the antenna with respect to the target. A maximum signal is received when the axis of the lobe passes through the target. The sensitivity of this system depends on the angular width of the lobe pattern. The operator adjusts the position of the antenna system for maximum received signal. If the signal strength changes appreciably when the antenna is rotated through a small angle, the accuracy with which the on-target position can be selected is great. Thus, in Figure 53-39, the relative signal strengths A and B have very little difference. If the energy were concentrated in a narrower beam, the difference would be greater and the accuracy better.

The remaining dimension necessary to locate completely an object in space can be expressed either as an angle of elevation or as an altitude. If one is known, the other can be calculated from one of the basic trigonometric functions. A method of determining the angle of elevation or the altitude is shown in Figure 53-40. The slant range (Figure 53-40A) is obtained from the radar scope indication as the range to the target. The angle of elevation is that of the radar antenna shown in part B of Figure 53-40. The altitude is equal to the slant range multiplied by the sine of the angle of elevation.

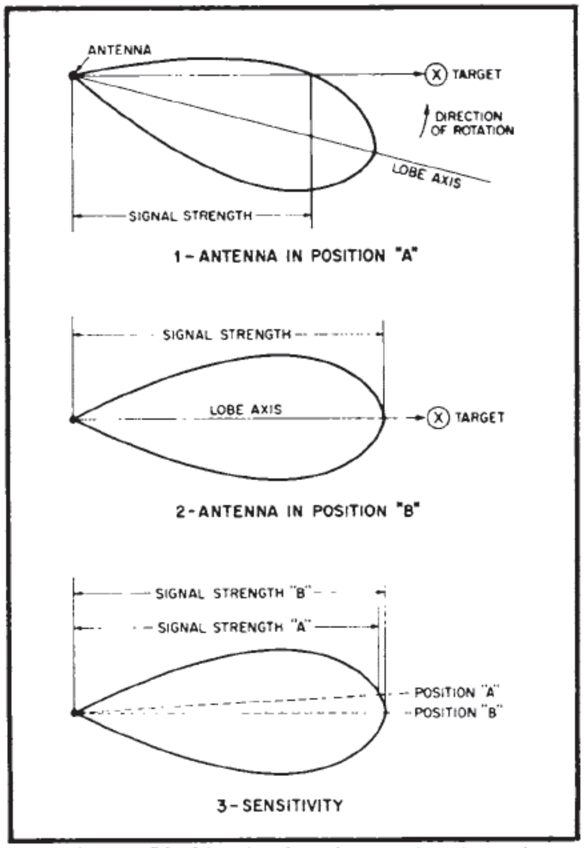


Figure 53-39 - Radar determination of azimuth or bearing.

In radar equipments with antennas that may be elevated, altitude determination by slant range is automatically computed electronically. In equipments (air search) where the antennas do not elevate, the altitude may be calculated by means of charts.

Search radars used for early warning nets do not require great precision in ranging or bearing, but do require the ability to locate targets at fairly long ranges. Therefore, they are normally designed with high power, wide beam angle, and fairly long pulse widths. Their target resolution (ability to accurately determine bearing and range) is not as good as that of radars used for another purpose such as fire control.

An air search radar detects and determines the bearing of aircraft targets at long ranges maintaining complete 360° surveillance from the surface to high altitudes. System constants must be chosen with this function in mind. Relatively low radar frequencies are chosen (P or L band 30-1000 mc) to permit long-range transmissions with minimum attenuation. Wide pulse widths (2 to 4 microseconds) and high peak power are used to aid in detecting small targets at great distances. Low pulse repetition rates are selected to permit greater maximum measurable range. Wide vertical beam width is used to ensure detection of targets from the surface to relatively high altitudes, and to compensate for the pitch and role of the ship. Medium horizontal beam width is employed to permit fairly accurate bearing resolution while maintaining 360° search coverage.

A surface search radar detects and determines the accurate range and bearing of surface targets while maintaining 360° surveillance for all surface targets within line-of-sight distance of the radar antenna.

Since the maximum range requirement of a surface search radar is primarily limited by the radar horizon, very high frequencies (X band) are employed to permit maximum reflection from small target reflecting areas, such as ship mast head structures and submarine periscopes. Narrow pulse widths (0.37 to 2 microseconds) are used to permit a high degree of range resolution at short ranges, and to achieve greater range accuracy. High pulse repetition rates (600 to 1000) are used to permit detection of small targets at line-of-sight distances. Wide vertical beam widths (10° to 30°) permit compensation for pitch and roll of the ship and to detect low flying aircraft. Narrow horizontal beam widths (1° to 3°) permit accurate bearing determination and good bearing resolution.

A radar system which indicates the bearing of targets must have some means of pointing its radiated energy in known direction. Practically all such radar systems accomplish this

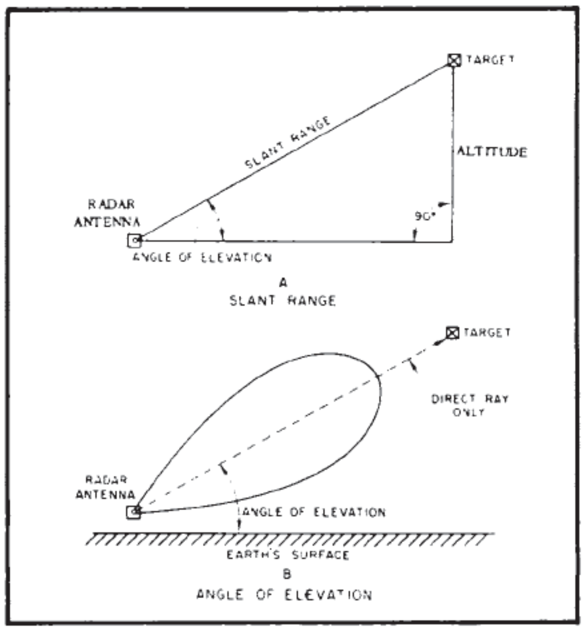


Figure 53-40 - Radar determination of altitude.

by constant 360° rotation of a motor driven energy transfer device such as an antenna, waveguide, reflector or director, or energy feedhorn.

Each antenna type has abilities to couple and project electromagnetic energy into space; also each has an ability to convert received energy into the forms that activate receiver equipments.

Experience shows that a parabolic dish, when properly adjusted focuses for projecting energy, will also serve at its best for accepting echo energy from space and directing it into the receiver system.

If the parabolic reflector is sufficiently large so the distance from any point within the dish to the focal point is several wavelengths, then QUASI-OPTICAL conditions exist and the emerging wave is a narrow beam. Sizes of reflectors, which are practicable for microwave work, have a diameter of 10 to 20 wavelengths to produce a beam width of approximately 5 degrees.

The quasi-optical theory is mentioned often in describing radar behavior. The word quasi means "similar" or "like." When you speak of microwaves from a high-frequency radar transmitter being quasi-optical waves in their behavior, you merely mean that invisible radar waves act like visible light waves. Therefore, in the make up of a radar antenna system that will adequately meet the requirements just discussed, it is possible to use (1) a central RADIATOR, or FEED, which compares with the bulb in a headlight (2) a PARABOLIC REFLECTOR, which compares with the silvered headlight cup, or reflector, and (3) a suitable means of positioning the beam in azimuth and elevation depending on the type of data desired.

The radiator can be either a dipole or a waveguide nozzle, commonly called a FEEDHORN, located at the focal point of the parabolic dish or reflector configuration. Waves radiating from this point strike the parabolic reflector in such a manner that they are reflected outward in a narrow beam similar to the beam from a headlight.

Figure 53-41 shows spherical wavefronts coming from radiator "A" at the focal point of a parabolic reflector. As the waves strike the reflector, they are straightened out and concentrated into a narrow beam of energy. The figure shows the spherical waves from the front of the radiator reinforcing the reflected waves as they travel outward from the reflector. However, this is not normally true of microwave antennas, for most parabolic reflector arrays have reflectors in front of the radiator. These reflectors concentrate most of the energy back toward the parabolic reflector, where it can be directed accurately, and thus prevent radiation of spherical unfocused waves in the forward plane.

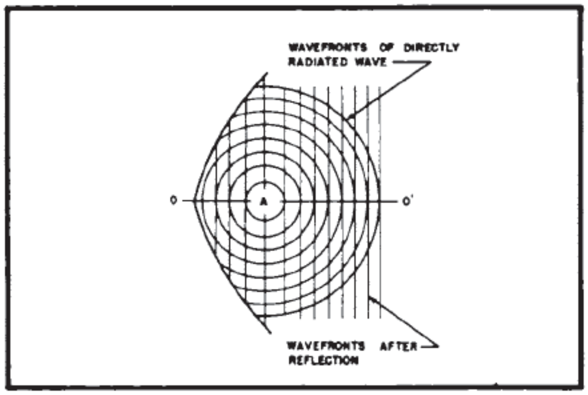


Figure 53-41 - Parabolic-reflected wavefronts.

Figure 53-42 shows the horizontal and vertical patterns produced by a typical parabolic reflector array. The reflected beams of RF energy can be made quite narrow; hence, the parabolic reflector array has great accuracy and high directivity.

The parabolic reflector can assume a variety of shapes, as can be seen in Figure 53-43. The shape of the reflector determines the shape of the RF beam. A perfect parabolic dish radiates a beam that is approximately the same angular width in both the horizontal and vertical plane. Other parabolic shapes have radiation patterns that are narrow in one plane and wide in the other. The TRUNCATED PARABALOID, shown in Figure 53-43, focuses the energy into a narrow beam in the horizontal plane. Since the reflector is truncated, or cut, so that it is shortened vertically, the beam spreads out vertically instead of being focused. Such a fan shaped beam is used to determine azimuth correctly.

The beam being wide vertically, will detect aircraft at different altitudes without changing the tilt angle of the antenna. The truncated parabaloid reflector can also be used in height

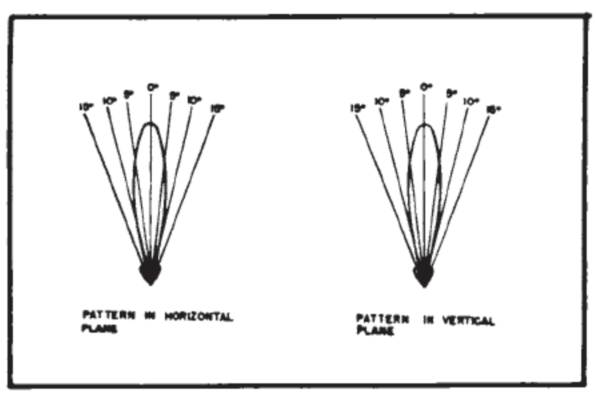


Figure 53-42 - Parabolic radiation pattern.

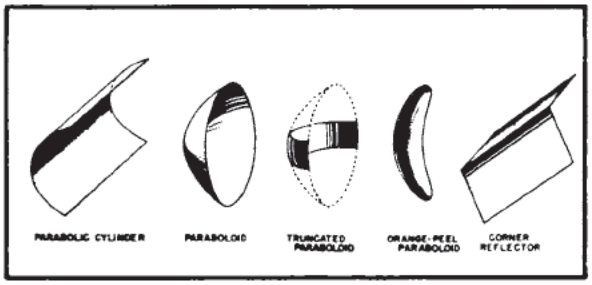


Figure 53-43 - Reflector shapes.

finding equipment. For an installation of this type, the reflector is mounted so that it is parabolic in the vertical plane and the energy will be focused in that plane. Since the reflector is thus shortened horizontally, the beam spreads out horizontally instead of being focused. Such a fan shaped beam is used to determine elevation accurately. Also, with such a wide beam in the horizontal plane, a greater area can be scanned in a shorter time than is possible with a narrow, pencil like beam.

The desired beam widths are provided for VHF radars by a broadside dipole array used with a flat reflector. When two or more half-wave elements placed one-half wavelength apart and parallel to each other are excited in phase, most of the radiation is broadside to the plane of the elements. A flat reflector, located approximately one-eighth wavelength behind the dipole elements, forms a unidirectional antenna system. An example broadside array is illustrated in Figure 53-44.

Besides the dipole radiator, which is used chiefly with radars having coaxial feeds, the nozzle feed is used frequently. The waveguide nozzle, or feedhorn, is merely an enlarged section of waveguide. It may be of several

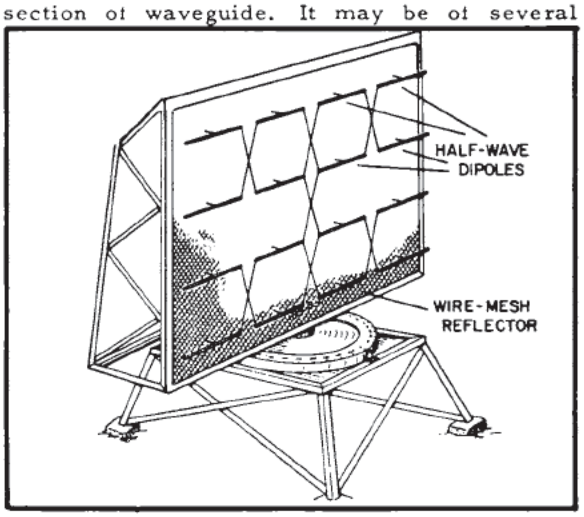


Figure 53-44 - Broadside array with flat reflector.

shapes, as illustrated in Figure 53-45. Horn radiators are adapted readily for use with waveguides, since they may serve not only to match the impedance of the waveguide to the external space, but also to produce directed wave patterns.

In most radars the feedhorn is covered with a window of polystyrene glass to prevent moisture or dirt from entering the open end of the waveguide.

In some radars that operate at extremely high frequencies, feed horns are used alone without a reflector. This is done only when the wavelength is shorter than the dimensions of the horn opening, because under those conditions the RF energy can be directed into the required narrow beam without excessive spreading.

Q28. Name the two main characteristics of a directional antenna.

Q29. Name two basic components that make up a radar antenna?

Q30. May reflector type antennas be designed by following optical principles? Why?

Q31. Name two types of radiating elements used with radar antennas.

Q32. What is the term used to describe the ability of radar equipment to distinguish small targets?

Q33. An antenna scanning through a sector in a vertical plane only allows a radar to determine what two factors about a target?

## 53-14. Antenna Safety Precautions

Radio frequency radiation from the antennas of high power radars present a definite hazard to personnel and equipment entering the beam path at short distances from the antenna.

Gasoline, oil, detonators, flares, or ammunition fitted with electrical primers should never be stored or permitted within a 50-foot radius of a high powered radar antenna. In the event such a condition must temporarily exist, every

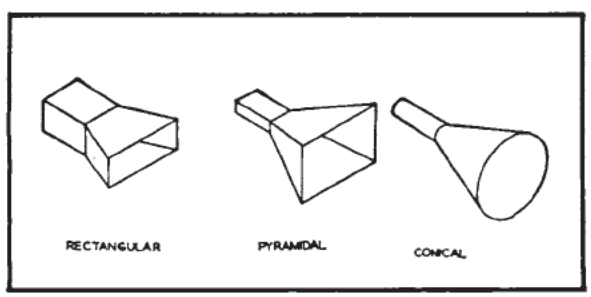


Figure 53-45 - Horn radiators.

A28. A directional antenna provides directivity and power gain.

A29. The two basic components that make up a radar antenna are the reflector and the radiating element.

A30. Yes. Reflector antennas exhibit optical characteristics.

A31. Two types of radiating elements used with radar antennas are dipoles and nozzles.

A32. Resolution.

A33. An antenna scanning through a sector in a vertical plane provides data from which range and altitude can be determined.

precaution should be taken to insure that the radar system is de-energized and remains so during this period.

Danger zones surrounding the antenna of radar systems, whose radiation is injurious to personnel, should be prominently marked to warn passers-by. Maintenance personnel should make certain that the high voltage is secured and measures taken to prevent it from being turned on by accident while work is being performed on the antenna.

Most radar antennas on shipboard are located high in the ship's superstructure or on a small platform on one of the masts. Also, many radar antennas are quite large and have a great amount of overhang beyond the pedestal or mounting base. Therefore, care should be taken to avoid being struck by a rotating antenna or being pushed off of a platform to the deck several feet below. All antennas that rotate in azimuth are equipped with safety switches so that maintenance personnel can de-energize the drive motor while working on the antenna assembly.

EXERCISE 53

1. What is the function of the magnetron in a radar transmitter?
2. Why are triodes unacceptable as high frequency oscillator circuits?
3. Basically describe the construction of a simple magnetron.
4. What is the difference between a negative resistance oscillator and an electron resonance oscillator?
5. What is the cycloid effect? To what does it pertain?
6. Describe the different types of anode blocks.
7. What controls the frequency of the magnetron?
8. What are "spokes?"
9. Describe the various methods of tuning?
10. Why is frequency stabilization of the magnetron desirable?
11. What is the purpose of an echo box?
12. What are some causes of magnetron arcing?
13. What is meant by the term "seasoning?"
14. What is the function of a duplexer?
15. Describe the construction of a duplexer, and how the TR and ATR function within the device.
16. Why is keep-alive voltage applied to a TR tube?
17. Why are switch transformers used with the TR and ATR tubes?
18. Why must caution be used when handling TR and ATR tubes?
19. Relate the terms "directivity," "horizontal beam," "vertical beam," and "bearing resolution."
20. What are lobes?
21. What is a parabolic reflector? What is a feedhorn? How are these two elements related?
22. Describe four types of reflectors, their characteristics, and their possible applications.
23. How can a pencil beam be formed?
24. What type of array would be used to obtain high resolution data?
25. Describe the safety factors to be considered when working on an antenna?
26. Define "working electron."
27. What is the phase relationship between adjacent anode segments in a magnetron operating in the pi mode?
28. From what source is the output energy of the magnetron derived?
29. At what potential is the anode of a magnetron operated?

# CHAPTER 54

# RADAR RECEIVERS

Of the thousands of watts peak power transmitted from a radar antenna, only a few milliwatts are returned to it as a target echo. The RADAR RECEIVER must be sensitive enough to detect this weak signal and amplify it to the level required for presentation.

This chapter deals with the study of the circuits used in the radar receiver and their purpose in receiving the target echo signals.

## 54-1. Basic Functional Block Diagram

Essentially the radar receiver is a special type of superheterodyne receiver. The basic functional block diagram is shown in Figure 54-1.

The function of the antenna is to convert the electromagnetic energy received from a target to electrical energy that is called the ECHO SIGNAL. By the use of a duplexer in the waveguide, the receiving antenna may also be used as the transmitting antenna.

The duplexer acts as a switching device. During the transmitting time, the highly sensitive receiver is protected by the waveguide design of the duplexer and its TR tube or tubes. Between transmitter pulses the duplexer connects the antenna to the input of the receiver.

The first block of the radar receiver is the mixer. The device used to provide mixing in a radar receiver is usually a crystal. The crystal is a non-linear device which mixes a continuous wave from the local oscillator with the incoming echo signal to produce an intermediate frequency possessing the same intelligence as the echo signal.

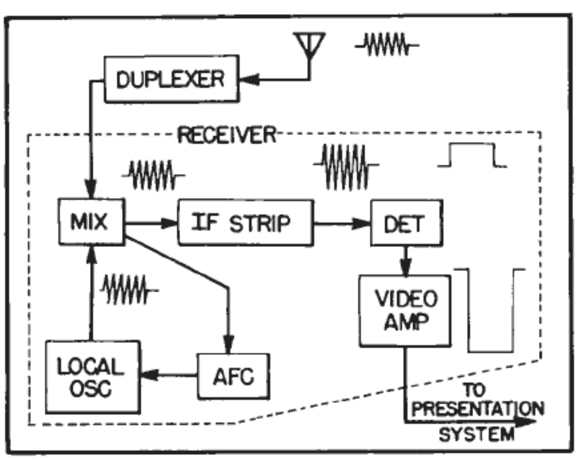


Figure 54-1 - Block diagram.

The local oscillator produces the continuous wave signal of the proper frequency. Its frequency is usually 30 Mc above the incoming frequency. After mixing action has occurred, the 30 Mc frequency difference which results will be the intermediate frequency. The local oscillator must be able to generate very high frequencies and provide some means of adjusting its frequency by electrical means.

After heterodyning, the signal output from the mixer is of a low amplitude. It must, therefore, undergo amplification. Since both the frequency and amplitude of the input pulse are critical, the amplifiers assigned to perform the function of amplification must do so contributing little or no distortion in the form of noise, cliping, or limited frequency bandpass response. To do this the amplifiers chosen will be of the tuned variety. A tuned amplifier can be adjusted to amplify only a narrow range of frequencies while rejecting all others. To provide a high gain, several IF amplifiers are connected in cascade. When this is done, the series of amplifiers is called the IF STRIP. The IF strip will substantially increase the amplitude of the signal. The IF strip has the primary control of receiver gain.

The mixer stage is considered the first detector. The second detector is placed after the IF strip to remove the carrier. The output of the second detector is a square wave pulse of voltage.

The output of the second detector is then applied to the input of a wide band amplifier called a VIDEO AMPLIFIER. The characteristics of a good video amplifier are a wide bandpass and adequate gain. Since the pulse of intelligence resembles a square wave, it will contain a large number of odd harmonics. To avoid distortion of the pulse, the video amplifier must amplify all of the harmonic frequencies equally. The output of the video amplifier is then sent to the presentation system.

Figure 54-1 shows a block designated as AFC.

This is the automatic frequency control circuit. The AFC provides a control voltage for the local oscillator. This control voltage adjusts the local oscillator to keep the difference frequency or IF constant.

The AFC circuit develops its control signal from an amplified and discriminated sampling of the IF signal produced by the AFC mixer assembly.

## 54-2. Velocity Modulation

To meet the requirements of the local oscillator in a radar receiver a new type of tube was developed called the VELOCITY MODULATED TUBE. Its principles of operation are related to those of the magnetron.

The ordinary electron tubes are DENSITY MODULATED. With a higher positive plate potential applied to an ordinary electron tube, the plate current will increase. This is due to the greater number of electrons attracted to the plate from the cathode space charge. The control grid of a triode tube varies the density of the electron stream between the cathode and the plate. The control grid has but little effect on the velocity of the electrons. Should the velocity of some of these electrons be changed, electrons would arrive at the plate with more or less kinetic energy.

An electron moving against an electrostatic field can be caused to increase its velocity. Increasing its velocity, the electron acquires more energy. The source of this new found energy is the electrostatic or DC field. If the electrons velocity is decreased, it will release energy. Its lost energy will be released to the AC or RF field which is developed by resonant circuits connected to grids within the tube. Therefore, there is an exchange of energy taking place between the electron and the electrostatic fields every time electron velocity is caused to change.

The circuit in Figure 54-2 is a basic velocity modulated tube. Its basic construction consists of an indirectly heated cathode, an accelerator grid, a set of buncher grids connected to a resonant cavity, and a collector plate. The accelerator grid, buncher grids and the collector plate are all made highly positive with respect to the cathode. The cavity resonator will act like a parallel resonant circuit.

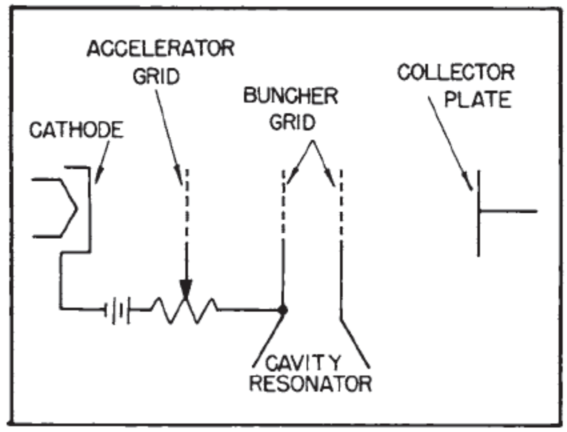


Figure 54-2 - Basic velocity modulated tube.

Figure 54-3 shows the cavity replaced by a parallel LC circuit. After electron emission has begun, electrons will leave the cathode at a high velocity. Some of these electrons will strike the accelerator grid and the buncher grids on their way to the highly positive collector plate. The electrons striking these grids will cause a small amount of current to flow in the accelerator and buncher grid circuits. In the buncher circuit, the current flow will cause the LC circuit to produce the flywheel effect. Figure 54-3 also shows the distance between the buncher grids as one-half wavelength. When the flywheel effect of the LC network commences, a sine wave of voltage will be developed across the buncher grids. When grid A is positive, grid B will be negative. When grid A is negative, grid B will be positive. There will be an electrostatic field existing between the grids in the

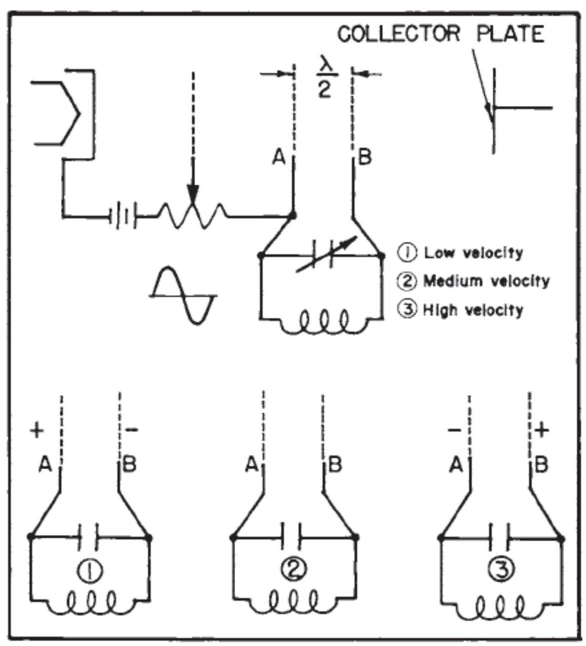


Figure 54-3 - Velocity modulated tube.

direction from the positive and toward the negative.

For the analysis of the buncher grid action on the electron stream moving from cathode to collector plate, three voltage conditions for the buncher grids will be used. The velocity of three different electrons will be compared with grid A more positive than grid B with grids A and B at equal potential, and with grid A less positive than grid B. The potentials will alternately be applied to the buncher grid by the flywheel effect of the tank circuit. Effectively there will be a sine wave of voltage applied across the buncher grids. When grid A is positive with respect to grid B, there will be an electrostatic field between the grids in the direction from A to B. An electron designated as #1 (Electron #1) will be accelerated to a given velocity as it passes the accelerator grid. As it nears the buncher field it comes under the influence of the field existing between the buncher grids. It is known that when an electron is caused to move with the direction of an electric field, the electron decelerates. Electron #1 will be decelerated. When the electron decelerates it loses part of its energy to the electrostatic (AC) field. Electron #1 will pass through the buncher grid B with less velocity than it had when it approached.

Assume that another electron designated as #2 approaches the buncher grids when there is no difference in potential between them. Electron #2 will pass through the buncher grids with the same velocity that it possessed before it arrived at the buncher grids.

Assume that another electron, #3, arrives near the buncher grid when grid A is negative with respect to grid B. The direction of the electrostatic field between the grids will be in a direction from B to A. Electron #3 approaching the buncher grids will come under the influence of this field. Since the electron will be moving in the opposite direction to the field, it will be accelerated. Therefore, the alternating voltage across the buncher grids causes the velocity of the electrons to vary. This is velocity modulation.

Figure 54-3 shows the electrons possessing different velocities. The analysis was begun with the decelerated electron #1 or the low velocity electron. Then followed #2 the medium velocity electron, and electron #3 the high velocity electron. Since they are each traveling at different velocities, given sufficient space, #2 will eventually catch up to #1. In fact, it is feasible that the high velocity electron will catch up to both the medium and low velocity electrons. When this occurs, all of the electrons will be travelling away from the vicinity of the buncher grids in a group. This grouping action is known as BUNCHING. A diagram showing this effect is illustrated in Figure 54-4.

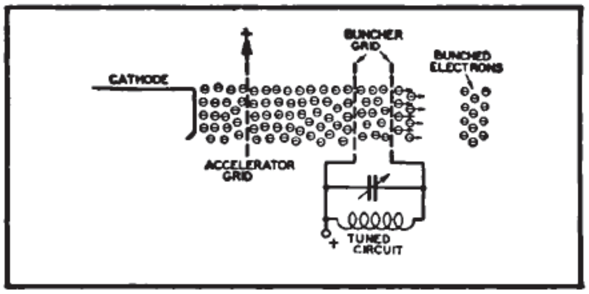


Figure 54-4 - Bunching.

The bunching action occurs between the buncher grids and the collector plate (this area is the drift space) at a point where the velocities of the electrons cause the paths of the electrons to converge. After that point, the electrons will move away at different velocities (near their original accelerated or decelerated velocities) toward the collector plate. The collector plate is usually located close to the point of electron divergence. Therefore, it may be said that the electron stream under the influence of the RF voltage applied to the buncher grids, moves down the tube in bunches rather than a continuous stream.

Q1. Name a difference between density modulation and velocity modulation?

Q2. What causes bunching action?

## 54-3. Basic Klystron

The velocity modulated tube described in Section 54-2 served no useful purpose. There was an exchange of energy between the electrons of different velocity and the RF electrostatic field, but in no way was this energy used with the possible exception of maintaining the oscillations of the cavity resonator. Even though the electron stream may be made to arrive at the collector plate in bunches, the effect of the bunching action differs in no way from the effect produced in a density modulated tube operated class "C" where current pulses reach the plate.

A way must be found for extracting the energy from the accelerating and decelerating electrons to serve a useful purpose. This can be accomplished by introducing a new set of grids into the tube. These new grids are called CATCHER GRIDS and are connected to a cavity resonator. This new device is called a basic POWER KLYSTRON and is shown in Figure 54-5.

For the sake of simplicity, the cavity resonator is replaced by its LC equivalent. The function of the catcher grids is to absorb any energy released from the electrons, in this case the bunched electrons. The tube is designed so that the bunching action of the elec-

A1. In density modulation, the velocity of the electrons passing from cathode to plate is constant.

A2. The RF voltage developed by the cavity resonator which is then applied to the buncher grids.

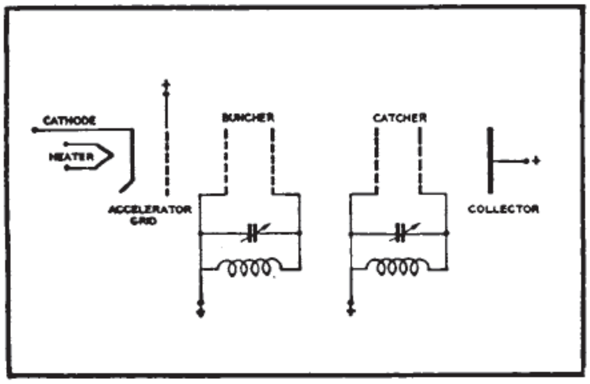


Figure 54-5 - Basic klystron.

trons occurs at a point equal to the midpoint between the catcher grids.

If the relative catcher grid potentials are as shown in Figure 54-6A, when each bunch of electrons reaches the first grid of this set, the AC field is such that it slows them down and thus absorbes energy from them. By the time the electron bunch reaches the second grid of the set, the relative potentials are reversed as shown in Figure 54-6B, because they are separated by a distance equal to a half wavelength. Therefore it takes the group of electrons approximately one half-cycle to go from one grid to the other. The second catcher grid also slows down the bunched electrons and absorbes energy from them. After delivering energy to the tuned circuit connected to the catcher grids, the spent electrons are removed by the positive collector plate. In this manner energy is removed from the passing electron bunches.

If a portion of the energy of the catcher tank is fed back in the proper phase to the buncher tank, the klystron can be caused to oscillate in much the same way as the TPTG oscillator. In this example, the catcher tank is analogous to

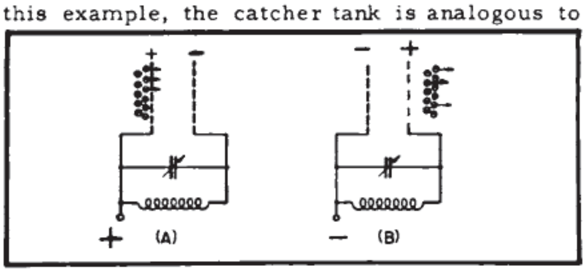


Figure 54-6 - Change of catcher grid potential.

the plate tank of the TPTG oscillator, and the buncher tank is similar to the grid tank.

The successful operation of the klystron as an oscillator requires that the energy needed for bunching be less than that delivered to the catcher. Amplifying action occurs in the klystron because the electrons pass through the buncher in a continuous stream and through the catcher in definite bunches.

Since a continuous stream of electrons enters the bunching grids, the number of electrons accelerated by the alternating field between the buncher grids on one half-cycle of oscillation is equaled exactly by the number of decelerated electrons on the other half cycle. Therefore, the net energy exchange between the electron stream and the buncher is zero over a complete cycle, except for the losses that occur in the tuned circuit of the buncher.

At the catcher a different situation exists. The electrons are traveling in bunches with the proper spacing so that they enter the catcher field only when the oscillating circuit is in its decelerating half-cycle. By this action more energy is delivered to the catcher than is taken from it. Thus the tube produces amplification.

The klystron may be used as an amplifier, oscillator, or mixer. For work at ultra high frequencies the tuned circuits of the buncher and catcher usually are cavity resonators as shown in Figure 5407. In this diagram, a grid is attached to each side of the cavity. These resonant cavities are very efficient, and are so small at extremely high frequencies that the entire cavity may be sealed inside the envelope of the tube. In this case, the cavity is tuned by varying the spacing of the cavity grids. Thus a slight flexing of the tube varies the effective capacitance of the tuned cavity circuit. In another type of construction, the grid connections are brought out through the envelope of the tube and an external cavity is used, clamped around the tube. In such a system, the cavity is tuned by changing its effective inductance. This can be done, for example, by screwing plugs into the periphery of the cavity. Energy may be coupled into or out of the cavity resonators by means of one-turn coupling loops, placed as shown in Figure 54-7, which provide coupling with the concentric magnetic flux within the cavity. Energy is carried from or to these loops by coaxial lines.

Q3. How is energy absorbed from the passing electrons by both of the catcher grids?

## 54-4. Reflex Klystron

When the basic klystron is used as an oscillator, it is critical to adjust. For that reason, the REFLEX KLYSTRON was developed. The reflex klystron differs from the basic klystron

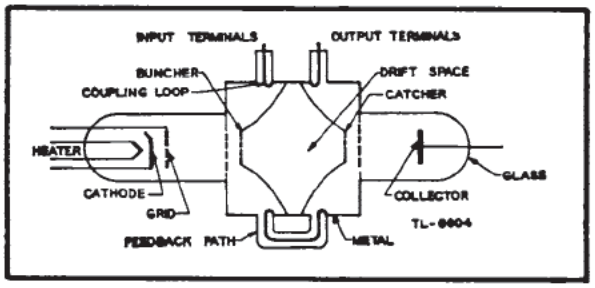


Figure 54-7 - Klystron tube with resonator.

discussed in Section 54-3 in that the same grids are used for both bunching and catching. Also the collector plate is replaced by a REPELLER PLATE. The potential applied to the repeller plate is negative. Electrons moving toward this plate will be repelled back in the direction of their origin. The repeller plate is the most negative element in the tube. A diagram showing the reflex klystron is shown in Figure 54-8.

The reflex klystron is similar in operation to the basic klystron. Electrons accelerated by the accelerator grid will be velocity modulated as they pass through the cavity grids. The cavity grids in the reflex klystron perform this function in the same manner in which it was performed in the basic klystron. The electrons after passing through the cavity grids will move at different velocities. Since the repeller plate is made highly negative, the electrons progressing toward it will stop and reverse their direction. The high velocity electrons will come physically closer to the repeller plate than either the medium or low velocity electrons. After repulsion, they will be directed back toward the cavity grids. In the reflex klystron, bunching action occurs on the return trip of the electrons. In fact, bunching occurs immediately before the electrons come under the influence of the RF field about the cavity grids. The distance that the electrons move before they are repelled by the negative repeller plate is a function of the

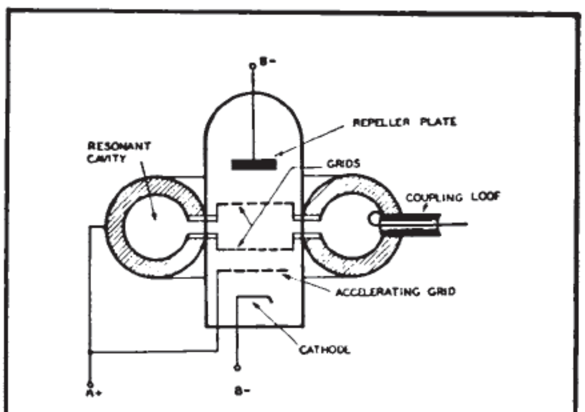


Figure 54-8 - Reflex klystron.

voltage values of the accelerating grid, the dc value of the voltage applied to the cavity grids, the dc voltage applied to the repeller plate, and the magnitude of the RF voltage coupled to the cavity grids by the cavity resonator. The voltages applied and the physical construction of the klystron should be of such values that the electrons will return to the cavity grids in bunches.

The potential of the cavity grids when the repelled electrons return is important. The bunched electrons should be returned when the potential applied to the cavity grids is such that the energy of the returning bunches will be absorbed. The maximum absorption of energy will occur when the bunched electrons reach the midpoint between the cavity grids in coincidence with the maximum positive peak of RF voltage between these grids. As the electron bunch reaches the midpoint the grid nearest the repeller plate must be positive in relation to the other buncher grid for correct alignment of the electrostatic field. The electron bunch will be decelerated in this field, thus expending some of its energy in sustaining RF oscillations within the grid cavity. Under these conditions, electrons leaving the cathode will receive maximum acceleration from the cavity field, while returning electron bunches will receive maximum deceleration. If the grids are seperated by approximately one half a wavelength, the electron bunch would pass through the first grid (one nearest the repeller plate) as its RF potential is zero and changing from negative to positive. The electron bunch would pass through the second grid when its potential is zero and is changing from negative to positive. After the returning electron bunches have given their energy to the cavity, they are absorbed by the cavity grid nearest the cathode and are returned to the power supply.

The cavity grids perform a dual function-velocity modulation and that of a catcher grid. The output from the tube is taken by use of the coupling loop shown in the diagram.

By proper adjustment of the negative voltage applied to the repeller plate, the electrons which have passed through the bunching field may be made to pass through the resonator again at the proper time to deliver energy to this circuit. Thus the feedback needed to produce oscillations is obtained and the tube construction is greatly simplified. Spent electrons are removed from the tube by the positive accelerator grid or by the grids of the resonator. The operating frequency of the tube can be varied over a small range by changing the voltage on the repeller plate. This potential determines the transit time of the electrons between their first and second passages through the resonator. However, the output power of the oscillator is affected considerably more than the frequency by changes in the magnitude of the repeller voltage. This is because

A3. It occurs because the catcher grids set up a decelerating field, and a decelerating electron gives up energy.

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the output power depends upon the fact that the electrons are bunched at exactly the decelerating half cycle of oscillating grid voltage. The volume of the resonant cavity is changed to change the oscillator frequency. The repeller voltage may be varied over a narrow range to provide minor adjustments in frequency.

Q4. What is the advantage of the reflex klystron over the two cavities, or the velocity modulated tube using both the buncher grids and the catcher grids.

54-5. Operational Analysis

It was mentioned that the electron bunches should arrive at the grids midpoint when the RF swing is at its maximum positive value on the grid closest to the repeller plate. It is not necessary for the electron bunches to return on the first positive half cycle. They may be returned on the second, third, or fourth positive half cycles. The positive half cycle in which the electrons are returned and bunching occurs determines the MODE OF OPERATION. Therefore, the mode of operation is determined by the transit time of the electrons. Transit time here means the time between which electrons leave the bunching grids and the time when the bunches deliver their energy to the cavity grids. Figure 54-9 shows the electrons being returned for the different operational modes. For the first mode, the bunching should occur 3/4 of a cycle after the average velocity electrons leave the bunching grids, the second mode of operation occurs 1 and 3/4 cycles after the average velocity electrons leave the bunching grids, the third after 2 and 3/4 cycles, and the fourth after

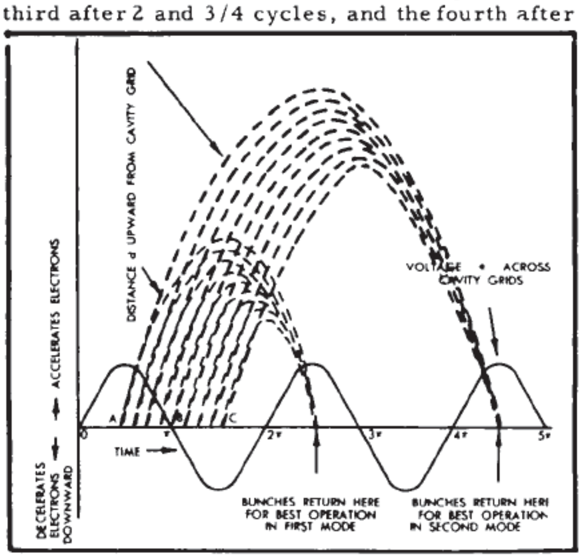


Figure 54-9 - Modes of operation.

3 and 3/4 cycles. In practical operation, either the second, third, or fourth modes are used.

The mode of operation is determined by the transit time of the electrons. The transit time is a function of both the accelerator voltage and the repeller plate voltage. Since the accelerating voltage is a fixed quantity, the mode of operation is controlled by the repeller plate voltage.

Figure 54-10 shows power output and frequency of oscillations as functions of the repeller voltage for three modes of operation. Notice that the frequency at the point of maximum output is the same for all three modes and is the resonant frequency of the cavity. In addition, note that the power outputs for the various modes at the resonant frequency are not the same and that the output is least at the highest mode. This can be explained by examining the factors which limit the amplitude of oscillations and which, in turn, limit the power output.

Power and amplitude limitations are due to overbunching as well as the usual losses in the oscillatory circuit. Overbunching occurs in the following way. As oscillations build up and the voltage on the cavities becomes greater, the amount of acceleration and deceleration increases. This causes bunching to occur in a shorter period of time, that is, in a time before the electrons reach the grids on the return trip. This tends to reduce the magnitude of the oscillations. In the higher modes of oscillations where the bunches are formed more slowly, the electrons are more susceptible to overbunching. The magnitude of the RF voltage that results from over-bunching is therefore lower, and oscillations are limited by this action to a lower amplitude than in the lower modes of operation.

As shown in Figure 54-10, the frequency of oscillations in a reflex klystron is variable to a limited degree in any of the modes of operation by varying the repeller plate voltage. When the repeller plate voltage is varied, it causes a bunch to return either a little sooner or a little later than normal. Off resonance, the amplitude of oscillations decreases by an amount determined by the Q of the cavity. In this tube the tuning range is small in comparison with the frequency of oscillations and varies somewhat from one mode to another. It is greatest in the highest mode, because bunching and debunching take place at a slower rate and because greater variation from the ideal time of return is possible without debunching. This would cause the amplitude of oscillations to drop below the usable output level.

Another perspective is that in the highest mode the interval between electrons leaving the grids and returning is greater. Therefore, the change in period represented by a given change

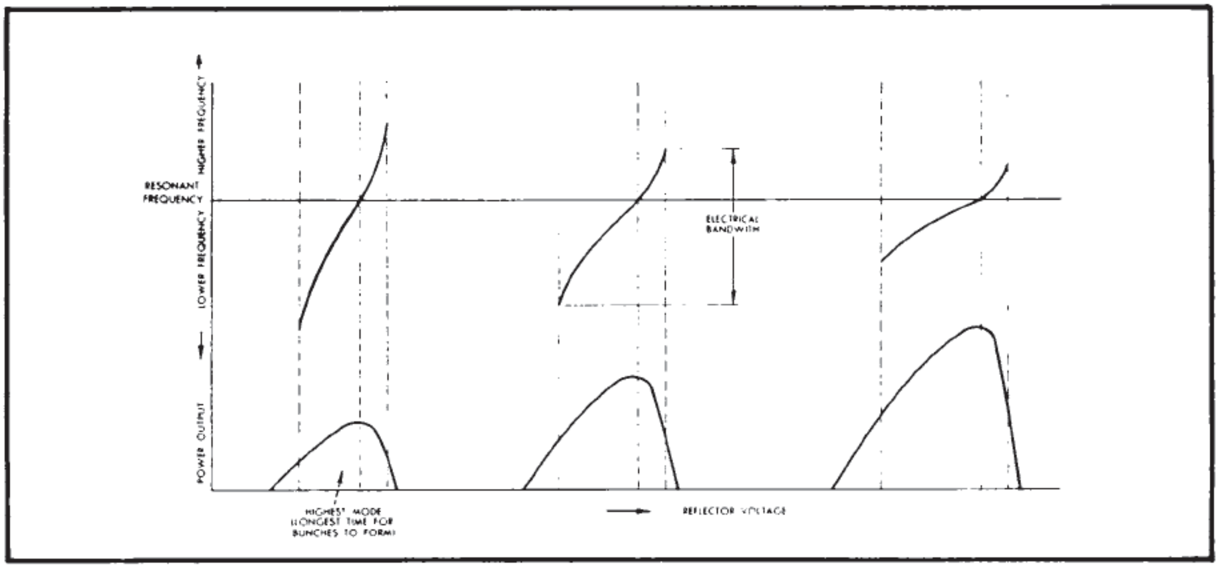


Figure 54-10 - Power output and frequency.

in frequency is a smaller portion of the interval. To illustrate, in the third mode the interval before return must be about two and three-fourths cycles. A small change in the period of the RF applied to the grids would therefore be only 3/11 as great a portion of the interval as it would if operation were in the first mode where the ideal time is three-fourths of a cycle.

The band of frequencies which can be obtained by varying the repeller plate voltage lies between the half power points shown in Figure 54-10. This range of frequencies is known as the ELECTRICAL BANDWIDTH. The output curves of the bandwidth are not symmetrical about the maximum output points. This results from the fact that if the repeller voltage is increased, not only does the bunching voltage decrease and cause the bunches to form at a later time, but the repeller voltage causes a quicker return. The effects of the two actions add to cause poor bunching at the time the electrons return, resulting in a rapid drop in output on the high side of the hump. At lower voltages, however, even though the bunching voltage decreases and causes slower bunching, the decreased repeller voltage causes a later return to the grids. In this way the two effects are counteracting and a greater change in repeller voltage is possible before the output drops below the usable level.

The choice of the point and mode of operation is a compromise among several factors. To begin with, there are three or four modes that have the necessary power output. Consequently it would appear that the correct choice would be the highest mode, for it gives the largest tuning range. The highest mode, however, is too sensitive to a change in voltage to be very well regulated. A change of one volt may cause a change of 0.5 Mc in the 3000 Mc oscillator. Since the power output modes are not symmetrical, the point of operation is usually chosen a little below the point of maximum output. This makes possible the tuning above the operating frequency by a greater amount than if the maximum point were used.

In practice, the reflex klystron is usually used in conjunction with an automatic frequency control circuit. This circuit controls the repeller voltage in such a way as to maintain the correct intermediate frequency. The automatic frequency control is usually provided by a frequency discriminator. Keep in mind that the frequency of oscillations is primarily determined by the dimensions of the cavity and that the repeller voltage is effective in making small changes in the frequency. Hence, in most reflex klystrons there is a coarse frequency adjustment that varies the cavity size in some way. The repeller voltage is the fine frequency adjustment.

Table 54-1 shows some of the operating characteristics of reflex klystrons. The data in this table gives some idea of the order of magnitude of the tube operating parameters. There is a wide variation between different tubes and different conditions of operation.

The K417 reflex klystron is one of the earlier types that was used for 10-cm operation. One feature of this tube was that in its early application it did not have a provision for controlling the frequency through a change in the repeller voltage since both the coarse and fine frequency controls changed the cavity grid spacing.

A4. When used as an oscillator, the reflex klystron is easier to tune. The tuning of the two cavity tube is accomplished by the tuning of both of the cavities which are interdependent on one another.

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Another 10-cm tube is the 707A (McNally) tube shown in Figure 54-11. In it the cavities are external to the tube and are not evacuated. This makes them susceptible to changes in temperature which results in changes in frequency. To establish good frequency stability, it is necessary to control the cavity temperature. The coarse frequency control consists of plugs which, when screwed into or out of the cavity, change its size. Fine frequency control is accomplished by the variable repeller plate voltage control.

The Shepherd-Pierce tube shown in Figure 54-12 is an all metal tube which is available for both 10-cm and 3-cm operation. The cavities are located inside the tube. Mechanical coarse tuning is accomplished through a strut on the side of the tube. The strut is adjusted by a screw which, in turn, varies the size of the cavity. The repeller voltage control serves as the fine frequency control. The 10-cm and 3-cm type Shepherd-Pierce tubes differ in the shape of the cavity and in the method of coupling the output.

Q5. What controls the frequency of the reflex klystron?

Q6. What advantage has the reflex klystron compared to the two cavity velocity modulated tube?

Q7. Why is the highest mode of operation in the reflex klystron not used?

## CRYSTAL MIXERS

### 54-6. Frequency Considerations

The purpose of a mixer in any receiver is to accomplish the process of heterodyning. When two frequencies are beat together, there are other frequencies produced. There are the original frequencies, the sum of the frequencies, and the difference between the frequencies in the output. The difference frequency is then used as the intermediate frequency.

To accomplish mixing action either a vacuum tube or a solid state device such as a crystal is used. Both of these devices are non-linear, a requirement for a mixer.

Ordinarily, a vacuum tube is used at low frequencies such as those encountered at communications frequencies. At frequencies above 1000 mc the vacuum tube produces a high level of noise that cannot be tolerated at the input of the radar receiver. The input to a radar receiver is a very small signal. If there is a high level of noise generated in the mixer, the signal will be lost in the noise. At these high radar frequencies, CRYSTAL MIXERS are used. They satisfy the non-linear requirement for a mixer and produce much less noise than the vacuum tube. One other limitation of the vacuum tube when used at high frequencies is the effects due to the transit time. Use of the crystal minimizes this effect.

Either a germanium diode or a silicon diode may be used to perform the function of the mixer, however, the silicon diode is preferred. They are preferred because of low conversion losses, low noise, better frequency response, and their ability to withstand momentary overloads.

### 54-7. Crystal

The silicon diode should not be confused with the function of the crystals which utilize the piezoelectric effect. The function of the crystal

| No. | Name | Mfr. | Freq. (Mcps) | Acc. Voltage | Beam Cur. (Ma) | Repeller Voltage | Control Grid Voltage | Power Output (mw) | Elec. Tuning (Mcps) |
|---|---|---|---|---|---|---|---|---|---|
| K417 | Klystron | Sperry | 3000 | 300-600 | 5-30 | +50 to −500 | +5 to +50 | 150 | 5 |
| 707A | McNally | W. E. and Raytheon | 3000 | 250-325 | 25-35 | 0 to −250 | same as acc. | 75 | 30 |
| 726A | Shepherd-Pierce | W. E. | 3000 | 300 | 22 | −20 to −300 | same as acc. | 100 | 20 |
| 723A | Shepherd-Pierce | W. E. | 9400 | 300 | 18-25 | −20 to −300 | same as acc. (internal connection) | 20 | 45 |

Table 54-1 - Typical reflex klystron tubes.

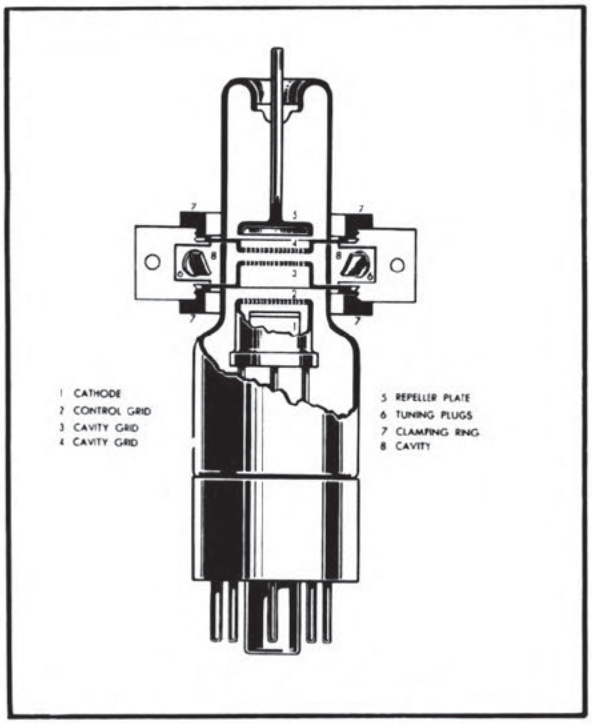


Figure 54-11 - 707A (McNally) tube.

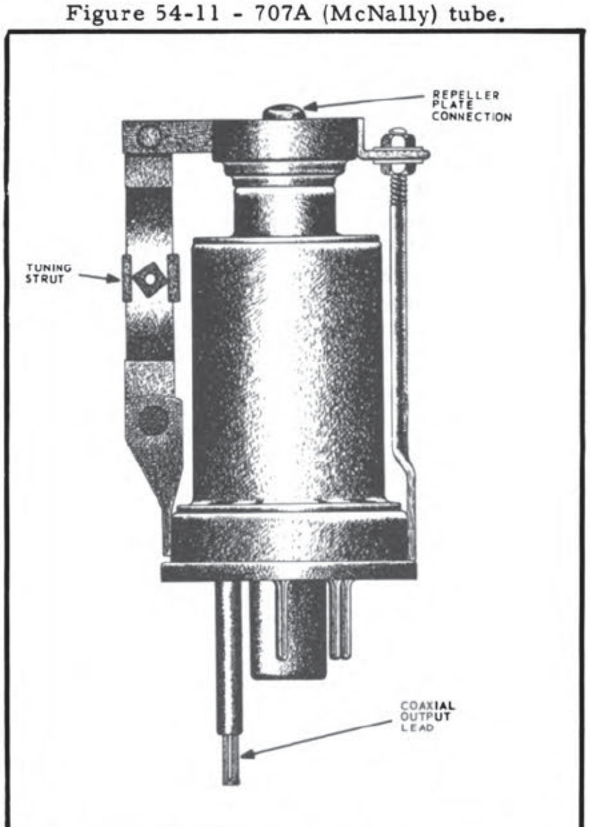


Figure 54-12 - Shepherd-Pierce reflex klystron.

in the silicon diode is quite different.

This silicon diode or semiconductor diode is called a POINT CONTACT DIODE. It is shown in Figure 54-13A. Unlike the junction diode, the point contact diode depends on the pressure or contact between a point and a semiconductor crystal for its operation.

One section consists of a small rectangular crystal of N- type silicon. A fine berylium-copper, bronze-phosphor, or tungsten wire called the catwhisker presses against the crystal and forms the other part of the diode. The reason for using the pointed wire instead of a flat metal plate is to produce a high intensity electric field at the point contact without using a large external source voltage. It is not possible to apply large voltages across the average semiconductors because of excessive heating.

The opposite end of the catwhisker is one of the terminals of the diode. It has a low resistance contact to the external circuit. Figure 54-13D illustrates a cutaway view of the point contact diode and the low resistance path to the external circuit. A flat metal plate on which the crystal is mounted forms the lower contact of the diode with the external circuit. Both contacts with the external circuit are low resistance contacts. The conventional symbol for the crystal diode is shown in Figure 54-13C. The arrow points in the direction of conventional current flow; electron flow is in the opposite direction, against the arrow.

During the manufacture of the point contact diode, a relatively large current is passed from the catwhisker to the silicon crystal. The result of this large current is the formation of a small region of P material around the crystal in the vicinity of the point contact, as shown in Figure 54-13B. Thus, there is a PN junction formed which behaves in the same way as the PN junctions previously described.

The characteristics of the point contact diode under forward and reverse bias are somewhat different from those of the junction diode. With forward bias the resistance of the point contact diode is higher than that of the junction diode. With reverse bias the current flow through a point contact diode is not as independent of the voltage applied to the crystal as it is in the junction diode. The point contact diode has an advantage over the junction diode in that the capacitance between the catwhisker and the crystal is less than the capacitance between the two sides of the junction diode. As such, the capacitive reactance existing across the point contact diode is higher and the capacitive current that will flow in the circuit at high frequencies is smaller.

Q8. What characteristic of the point contact diode permits its use as a rectifying device?

A5. The dimensions of the resonant cavity.

A6. The reflex klystron is easier to tune when it is used as an oscillator.

A7. Because at that mode, the frequency adjustment is critical. With a small change in voltage applied to the repeller plate a large change in frequency results.

A8. It offers more resistance in one direction to the flow of current, and less in the opposite direction. It is also a non-linear device.

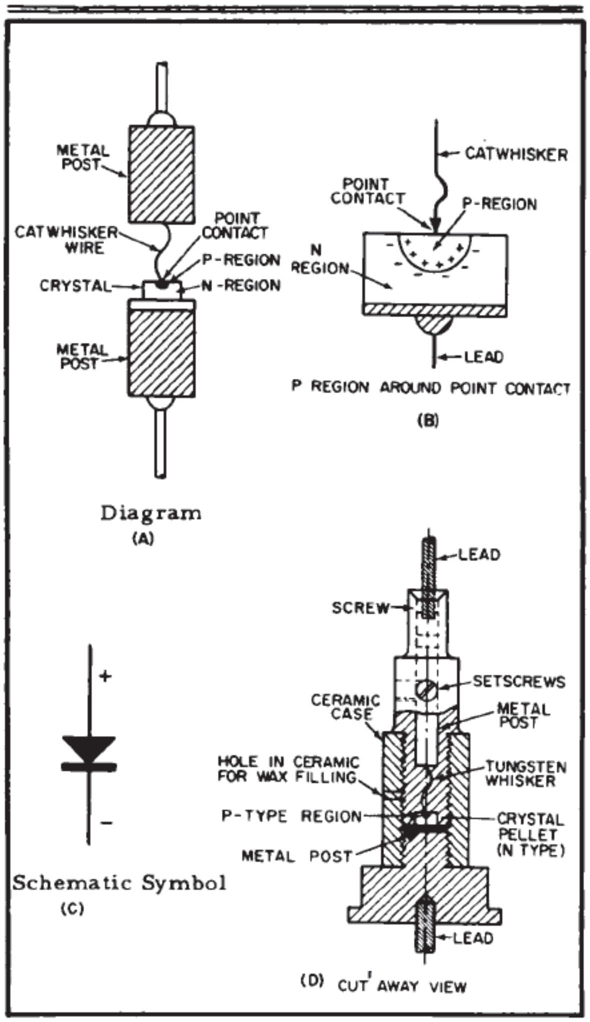


Figure 54-13 - Point contact crystal diode.

## 54-8. Single-Ended Crystal Mixer

A basic SINGLE-ENDED CRYSTAL MIXER is shown in Figure 54-14. The inputs to the circuit are the local oscillator continuous wave frequency, and the modulated RF input signal. The crystal will perform the process of rectification which will result in an average dc current that may be recorded by the dc current meter. The output of the crystal will be a pulsation of dc which will contain many harmonics plus the sum and difference frequencies produced by the heterodyning action of the non-linear crystal. The only frequency component of concern is the difference frequency which will become the modulated intermediate frequency. All of the other frequency components of the crystal output must be eliminated. This elimination is accomplished by the use of a filter network. The desired intermediate frequency will be developed across the resonant circuits tuned to the value of the intermediate frequency. Because of the flywheel effect of the tank circuits, a full modulated sine wave output will be produced.

The crystal current meter will indicate whether the tuned circuit is tuned to the proper frequency. It will also indicate the crystal condition. If there is zero current, the crystal could be open. If there is an excessive amount of current, the crystal could be shorted.

Q9. What is the indication on the dc crystal current meter when the I. F. tank circuit is tuned to the proper intermediate frequency? Why?

## 54-9. Balanced Crystal Mixers

One disadvantage of the single-ended crystal mixer is that it does not eliminate noise. This disadvantage is overcome by the use of a BALANCED CRYSTAL MIXER.

At radar frequencies, waveguides and coaxial lines are used to couple RF energy from one point in a circuit to another. In these applications, the mixer crystals are inserted directly into the waveguide, waveguide section, or coaxial line.

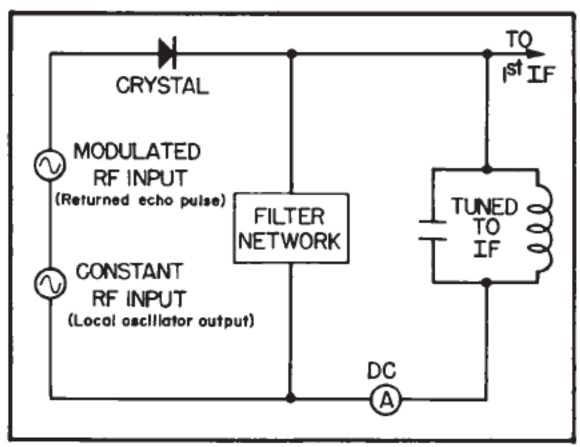


Figure 54-14 - Basic single-ended crystal mixer.

A balanced crystal mixer is shown in Figure 54-15. In this type of mixer, the crystals are inserted directly into each side of the coaxial section. The local oscillator input is applied from the probe as shown in the diagram. The local oscillator voltage will be applied to each crystal with the same magnitude and phase relationship.

The input for the echo signal or modulated RF input signal is on the left side of the section. The generator shown in that position represents the TR device. Crystal #1 is located 1/4 wavelength from the TR device. The second crystal #2 is located a half wavelength from crystal #1. Crystal #2 is located at 1/4 wavelength from the shorted right-hand side of the section. The voltage applied to crystal #2 will be a maximum because it is located 1/4 wave from the shorted end. The voltage at crystal #1 is also at maximum because it is located 1/4 wave from the TR device which may be assumed to be a short. Since crystal #1 is located a half wave away from crystal #2, there will be a difference of 180° between the echo voltages applied to them.

The output terminals for the circuit shown in the diagram will couple the IF signal from the crystals to the balanced transformer. Since there is a difference between the phase of the echo signals applied between the two crystals, and because the voltages applied to the crystals from the L.O. input probe are in phase, there will be a condition where both signals applied to crystal #1 will be in phase, and voltages applied to crystal #2 will be 180° out of phase. This means that an IF signal of one phase will be produced by crystal #1, and an IF signal of the opposite phase will be produced across crystal #2. When these two signals are applied to the balanced output transformer, they will add. If the outputs were of the same polarity, they would cancel across the balanced transformer.

It is this action that eliminates the noise. The noise components which are introduced with the L.O. probe, are applied in such a fashion so that they will be in phase with the voltages applied to each crystal. This means that they will be cancelled at the output transformer. If they were applied out of phase with the local oscillator signal, the noise would appear in the output.

In this example, coaxial sections of transmission line were used. For higher frequency applications, waveguide sections are used. Figure 54-16 shows a balanced mixer using waveguides. Since the waveguide section forms the letter T, the device is sometimes referred as the MAGIC T. Its operation is similar to the operation of the balanced mixer using the coaxial section. The difference between this type and the coaxial type is that the L.O. input and the RF input are not introduced into the waveguide by probes. They are introduced into the guide through the arms of the T section.

The position of both crystals are such that they are located 1/4 wavelength (at a point of maximum voltage) from their respective ends, and 1/2 wavelength from each other. The output is taken from the crystals and applied to a balanced transformer

The echo signal is applied through arm A and is distributed in arms C and D. There is a 180 degree phase shift in the phase of these two signals. This phase shift is shown by the direction of the arrows in both sides of the guide. The local oscillator input is applied to the guide through arm B. This input is also distributed equally in both arms of the guide. However, there is no phase shift between the L.O. component in each of the arms. These voltages are in phase. Therefore, at crystal #1, the signal input (echo signal) and the local oscillator signal are in phase. At crystal #2 the local oscil-

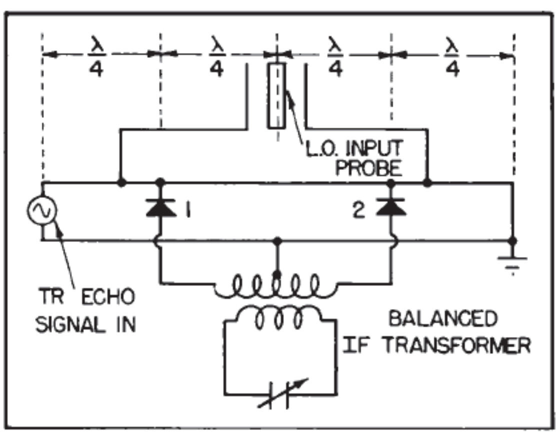


Figure 54-15 - Balanced crystal mixer (coaxial).

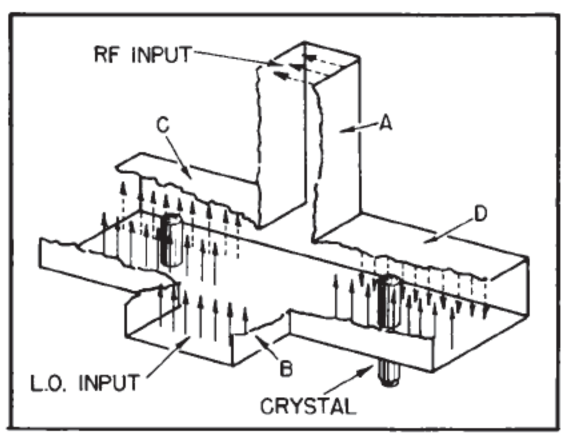


Figure 54-16 - Magic T or balanced hybrid mixer.

A9. Minimum current on the crystal current meter because at resonance, the tank circuit will offer a maximum impedance.

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lator signal and the echo signal are 180 degrees out of phase. The phase relationship of these two signals is shown in the diagram. The outputs from the crystals will then be applied to the balanced transformer. Since the signals are out of phase, the noise component which is in phase with the L. O. frequency will be cancelled in the same manner as it was in the coaxial section.

The difference between the balanced mixer using the coaxial section and the one using the hybrid junction is that the hybrid junction is used at higher frequencies such as the radar frequencies. The output of both of these types is then applied to an IF amplifier where the signal will be amplified.

Q10. How is noise eliminated in the balanced crystal mixer?

Q11. Why are the crystals located 1/4 wavelength from the ends of the devices—both the coaxial and the magic T?

## IF SECTION

### 54-10. Wide Band Amplifier

The IF amplifier in a normal radio receiver is a high gain tuned amplifier adjusted to amplify the modulated IF signals. They will normally pass a narrow band of frequencies. The demands made of the IF amplifier used in a radar receiver are different from those made of the ordinary IF amplifier used in conventional radio receivers. The difference is dictated by the nature of the signal to be amplified. Video pulses encountered in radar receivers are abundant in odd harmonics. The video pulses resemble square waves. If an amplifier is used that will not amplify all of the harmonic frequencies equally, the output waveform will be distorted and inaccurate range determination will result. Therefore, the bandpass of the IF amplifiers used in radar receivers will be considerably greater than the bandpass of IF amplifiers used in ordinary radio receivers. The IF amplifier in the radar receiver imposes the bandwidth limitations for the entire receiver, and controls the overall gain of the system.

The IF amplifiers used in radar receivers may be composed of more than one stage of amplification. All the IF stages taken as a group is normally refered to as the IF section or "IF strip." The IF section may be composed of IF stages which may be single tuned, stagger tuned, double tuned, or sometimes untuned. They may be connected cascode or cascade.

The IF section shown in Figure 54-17 is composed of conventional wide band amplifiers (pentodes) connected in cascade (plate to grid). Tuning of the amplifiers is accomplished by the variable inductors $L_1$, $L_2$, $L_3$ and $L_4$. These inductances associated with the total shunt capacitance of the stage serve as resonant circuits.

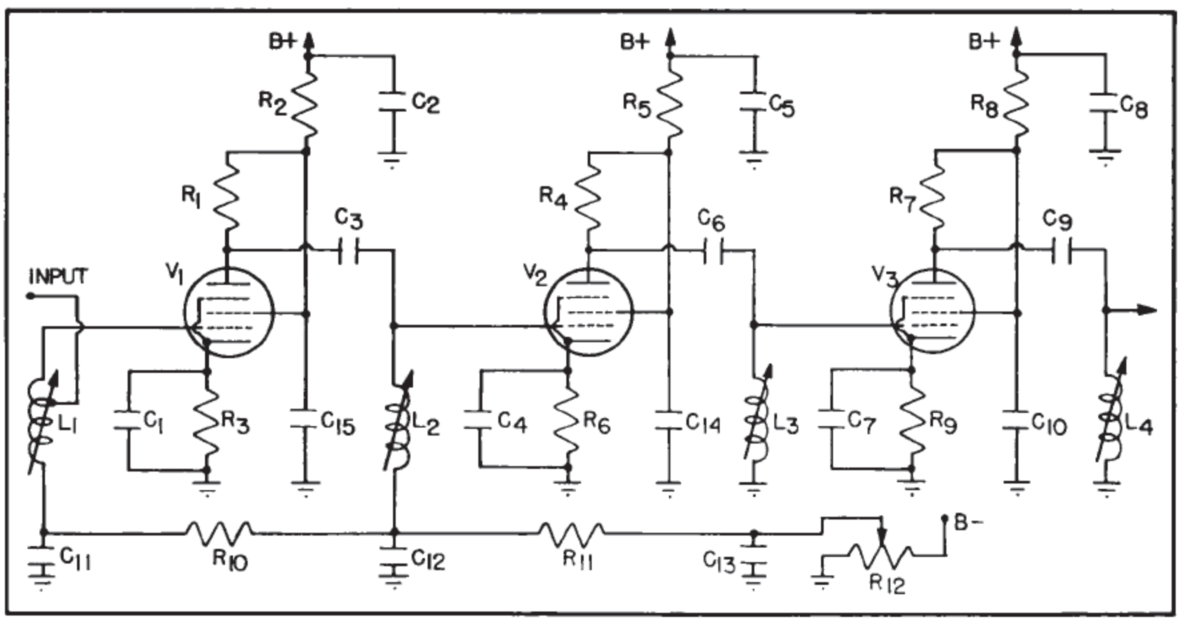


Figure 54-17 - Single tuned IF amplifier.

Each of these circuits are tuned to the IF frequency. The Q of the coils will help control the bandwidth of the amplifier.

Gain control is provided by the RC network composed of $C_{11}$, $C_{12}$, $C_{13}$, $R_{10}$, $R_{11}$, and $R_{12}$. The negative voltage provided from this network will vary the bias voltage on the grids of tubes $V_1$ and $V_2$. The function of the capacitors in this circuit is to bypass the RF to ground. Decoupling is also provided in this circuit by the bypass capacitors located in the plate circuit of each tube.

To determine the gain of an IF amplifier circuit, the gain per stage is calculated by a Norton's equivalent circuit as shown in Figure 54-18A.

In referring to the IF amplifier circuit of Figure 54-17, the Norton's equivalent circuit for a pentode wide band amplifier for any one stage results in the equivalent circuit illustrated in Figure 54-18A. Here gmeg, $r_p$ and output capacitance $C_o$, represents the plate circuit of the pentode tube of one stage. The input impedance to the next stage is represented by $C_i$ and $R_i$; where $R_i$ is the resistive loss between cathode and control grid of next stage. Resistor $R_d$ represents the coil dissipation and $C_d$ represents the sum of the wiring capacitance and the inter turn (sometimes referred to as distributed) capacitance of the coil. Combining the parallel resistors $r_p$, $R_L$, $R_d$ and $R_i$ into $R_{eq}$ is shown in Figure 54-18B. Also $C_{eq}$ is

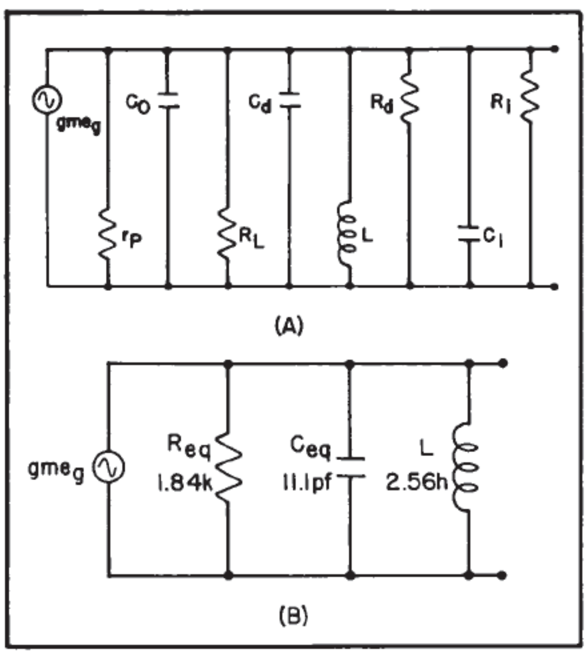


Figure 54-18 - Norton's equivalent circuit for a single stage.

illustrated which is the parallel equivalent of $C_o$, $C_d$ and $C_i$. Referring to Figure 54-17, utilizing a 6AK5 type pentode, the gm is 5100 micromhos and the resistor $R_L$ is 2 K ohms. The cathode resistor is 180 ohms and the decoupling resistors are adjusted to supply 120 volts for plate and screen grid. The characteristics of a 6AK5 tube, show the Q point to occur at a point where grid bias equals minus 2 volts and $r_p$ equals 500 K ohms.

Utilizing a common IF frequency of 30 Mc the value of $R_i$ equals 55 K ohms and $R_d$ equals 30 K ohms, approximately. In Figure 54-18B, the parallel combination of $r_p$, $R_L$, $R_d$, and $R_i$ equals $R_{eq}$ and 500 K, 2 K, 30 K and 55 K ohms in parallel equal the value of 1.84 K ohms. The output and input capacitances of the 6AK5 tube are 2.1 and 4 picofarads respectively. Therefore in Figure 54-18A, $C_o$ will have a value of 2.1 pf, $C_i$ of 4 pf and the distributed capacitance, $C_d$, is assumed to be 5 picofarads. The value of the total capacitance $C_{eq}$ in Figure 54-18B will equal 11.1 picofarads. Because the inductance must form a resonant circuit with $C_{eq}$ at 30 Mc:

using the resonant frequency formula

$$f_o = \frac{1}{2\pi\sqrt{LC}} \qquad (11\text{-}16)$$

and solving for L

$$L = \frac{1}{4\pi^2 f_o^2 C}$$

L will be adjusted to a value equal to 2.56 microhenrys.

The total capacitance is important because it together with the value of the inductance will determine the frequency to which the amplifier will be tuned. The values of the total resistance, and total capacitance will also control the value of the amplifier bandwidth and amplifier gain. The gain is determined by the following equation:

$$\text{Gain} = \text{gm Req} \qquad (54\text{-}1)$$

Where: Gain = mid-band amplification

gm = transconductance of the tube

Req = total resistance

Therefore: $A = 5100 \times 10^{-6} \times 1.84 \times 10^3$

$= 9.4$

In the equivalent circuit shown in Figure 54-18, the total resistance is shunted by the resonant circuit composed of L and C. At resonance, the

A10. The noise is in phase with the local oscillator signal, and will be cancelled at the balanced transformer.

A11. The 1/4 wavelength point is a point of maximum voltage.

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impedance of the network will be maximum. The total impedance that the tube "sees" will be less than the value of the total resistance. Since the ratio of the impedance of the LC network to the total resistance is so high, the impedance that the tube "sees" is assumed to be equal to the value of the shunting resistance. The gain thus computed is only for the first stage of amplification. If it is desired to determine the amplification of the entire strip, the gain of each stage will have to be considered.

The bandwidth of the amplifier is also controlled by the value of the resistance and capacitance. The equation for the determination of the bandwidth of the first stage of amplification is as follows.

$$BW = \frac{1}{2\pi \text{ Req Ceq.}} \qquad (54\text{-}2)$$

This formula should not be taken as the overall bandwidth formula but applicable to only one stage. The number of stages will affect the overall bandwidth.

The gain-bandwidth determination will give an indication of the merit of an amplifier. Their product (gain x bandwidth) will render a figure of merit that may be used to compare the merits of different IF amplifiers.

When several stages such as those just discussed are cascaded, the total mid-band amplification is the product of all of the mid-band amplification of all the stages. Therefore, the number of stages of amplification affects the overall mid-band amplification of an amplifier strip. The greater the number of amplifiers in a strip, the greater will be the total gain. Referring to the circuits of Figure 54-17, assuming there are six stages in cascade the gain is therefore $(9.4)^6$ or 930,000. There is one disadvantage to cascading many stages of amplification. When the number of stages is increased, the total bandwidth decreases. This decrease in bandwidth may be compensated for by stagger tuning successive stages of amplification.

The gain of an amplifier IF strip is very high. Gain value for radar IF strips in the region of a million are not uncommon. Because of this large gain, caution must be taken to prevent even the smallest amount of regenerative feedback. Even a very minute value of voltage of the proper phase fed back can cause the amplifier to break into oscillations. Because of this, great care is taken to eliminate the possibility of feedback. Extensive decoupling in the plate and filament circuits is employed. IF strips in radar sets are usually isolated and well shielded to prevent the interaction of any undesirable fields.

One method of avoiding the gain and bandwidth complications encountered in the single tuned amplifier is to use a double tuned IF amplifier One stage of such is shown in Figure 54-19. The reason this circuit is called a double tuned IF amplifier is because of the two resonant circuits used, one in the plate circuit and the other in the grid circuit of the following stage. The advantage of using a double tuned IF amplifier as compared to the single tuned IF amplifier is that the double tuned amplifier has a greater bandwidth and an improved frequency response.

Q12. What is indicated by a given value of gain x bandwidth?

## 54-11. IF Input Stage

The receiver system that is described in this chapter does not utilize any stages of RF amplification before the mixer stage. This means that the signal applied to the first IF amplifier will be very small. The possibility of a large amount of noise being generated under these conditions is very great. Remember, the amplification of an IF section may be as high as a million. This means that any noise generated in the first IF stage will also be amplified by that amount. To increase the signal-to-noise ratio, some stages of special amplification are used preceeding the normal IF input. These stages of additional amplification are called the IF INPUT STAGES or PREAMPLIFIER STAGES. One feature of these circuits compared to the normal IF section is that they are connected in cascode instead of cascade. The purpose of this

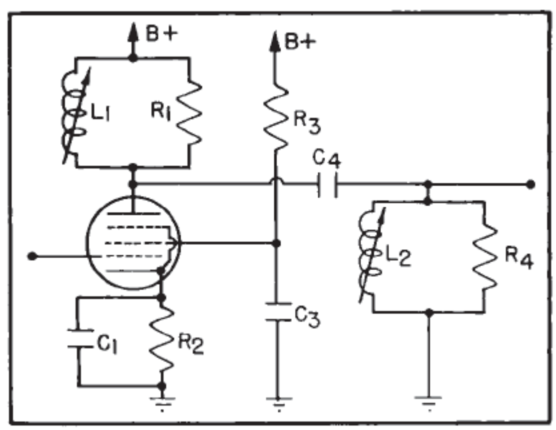


Figure 54-19 - Double tuned IF amplifier stage.

circuit arrangement is to provide a good signal-to-noise ratio. The cascode input stage is shown in Figure 54-20. This circuit provides approximately the same bandwidth and gain of a pentode with a noise figure equal to that of the first triode, $V_1$. Normally, the noise figure is established by the first amplifier stage. If its signal-to-noise ratio is low, the ratio of the entire receiver will be low.

$V_2$ is a grounded grid amplifier. The inductor $L_1$ in the input circuit is tuned to resonate at the mid-band frequency. Inductors $L_2$ and $L_4$ are likewise adjusted. Inductor $L_3$, $C_2$ and the grid to plate capacitance of $V_1$ form a resonant circuit which is also tuned to resonate at the mid-band frequency.

The voltage amplification of tube, $V_1$, is nearly equal to unity, but the amplification provided by $V_2$ is sufficiently high to provide a gain at the mid-band frequency nearly equal to that of a pentode.

For additional information concerning cascode amplifiers (grounded grid amplifiers), consult Chapters 38, 39 and 40.

Although the tuned circuits control the gain of the mid-frequency, they also account for some of the distortion present in a wide-band amplifier. The video pulse of voltage applied to the input of an IF strip is similar to that shown in Figure 54-21A. Since the rise time of the input pulse is short, and the inductive and capacitive components which form the resonant circuits cannot respond to the sudden changes of voltage; the output waveforms leading and trailing edges are distorted. This is called TRANSIENT DISTORTION. This distortion is shown in Figure 54-21B. This effect may be reduced by increasing the bandwidth of the resonant circuits which, in turn, increases the bandpass of the amplifier. One of the most common methods used to increase the bandwidth of the resonant circuits is by shunting their coils with a resistance. These resistors are called SWAMPING RESISTORS, and the process is known as SWAMPING. Since the methods used for compensating for distortion in wide band amplifiers is similar to the methods used to compensate in video amplifiers the methods of frequency compensation will be discussed when video amplifiers are considered in detail.

Tuning of the IF amplifier is a process seldom required. Since the tuning is accomplished at the time of manufacture, little adjustment is necessary. This, of course, excludes the necessity for adjustment when the components age and cause the response of the amplifier to change. Adjustment is sometimes necessary when the equipment is moved over a considerable distance, or when the equipment has stood idle over a considerable length of time. When adjustment is required, it may be accomplished through the use of instructions provided in the instruction manual for that particular type of equipment. These adjustments are critical, and should be made several times to insure accuracy. Particular care should be taken when the amplifier stages are stagger tuned.

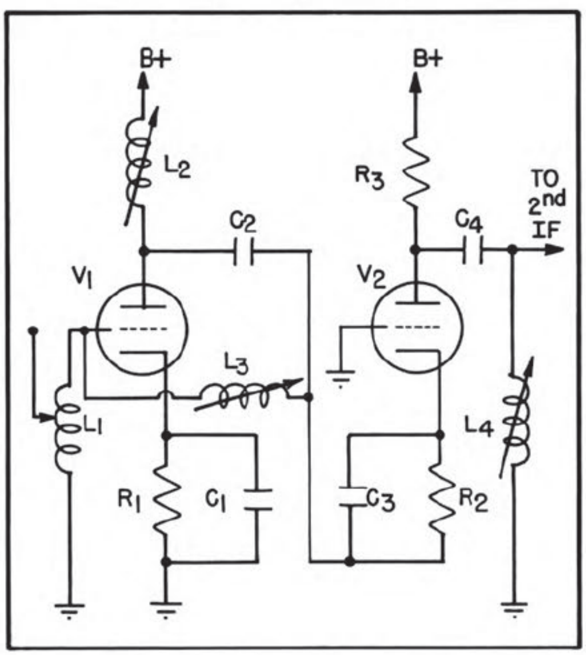


Figure 54-20 - IF input stage.

Q13. What is the purpose of cascoding the IF input stage?

Q14. What controls the signal-to-noise ratio of a radar receiver?

## 54-12. Detection

After the modulated video signal has been amplified by the IF strip, it is then applied to the detector. This stage is also called the second detector or the demodulator. It is called a demodulator because it separates the modulation from the RF carrier. One of the simplest

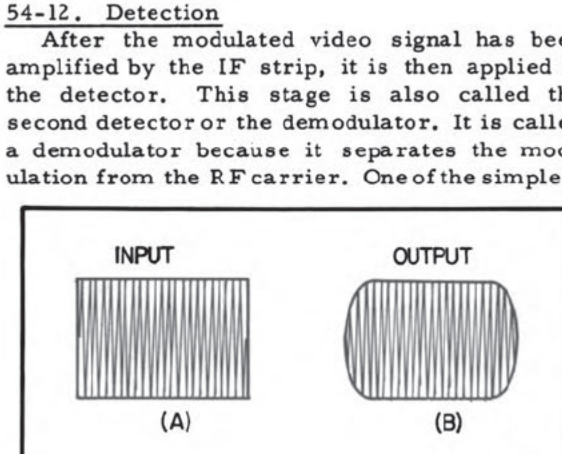


Figure 54-21 - Transient distortion.

A12. It is a figure of merit that will indicate the comparative worth of the circuit.

A13. To improve the signal-to-noise ratio.

A14. Ordinarily, the signal-to-noise ratio of the first stage.

methods of accomplishing demodulation is by use of the diode detector shown in Figure 54-22. Although more complex circuits are employed to affect detection, the simple circuit shown in Figure 54-22 will suffice for explanation. The input of the diode detector is tuned to respond to the mid-band frequency of the RF carrier. This circuit is nothing more than a rectifier. It will remove one half of the modulated signal. It will also, through the use of appropriate filters, remove the RF from the envelope. The output of the detector is then sent to the video amplifier.

## VIDEO AMPLIFIER

### 54-13. Frequency Response and Distortion

In both the wide-band amplifier and the VIDEO AMPLIFIER, there is a need to insure equal amplification over a wide range of frequencies. Video pulses are square waves abundant in odd harmonics. If all of these harmonics are not amplified equally, pulse distortion will result. The output waveform from an amplifier which does not equally amplify all of the harmonic content of the square wave pulse is shown in Figure 54-23. Figure 54-23A shows the type of output obtained from an amplifier with poor low frequency response. Figure 54-23B shows the output obtained from a circuit with poor high frequency response. It is conceivable that one amplifier circuit could give both poor high and low frequency response. This type of amplifier is shown in Figure 54-24. It is an RC coupled amplifier. To examine the limitations in frequency response of the pentode RC coupled amplifier in Figure 54-24, it is necessary to

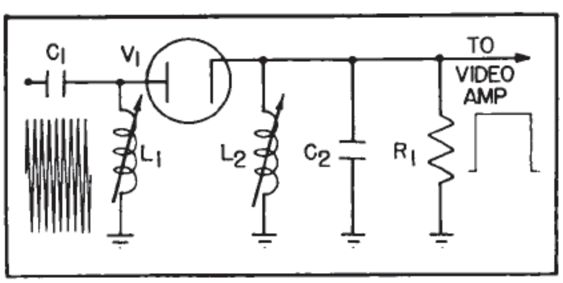


Figure 54-22 - Diode detector.

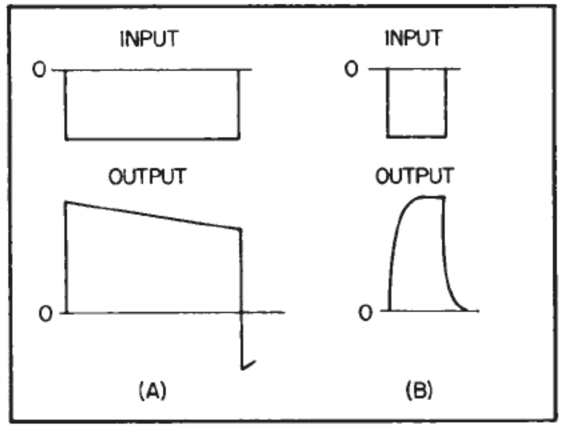


Figure 54-23 - High and low frequency distortion.

construct the equivalent circuit for the amplifier stage. Figure 54-25 is the basic equivalent circuit, the low frequency equivalent circuit, and the high frequency equivalent circuit. The medium frequency equivalent circuit will not be shown. If it is desired to review the equivalent circuits for RC coupled pentode amplifiers, refer to Chapter 20.

Since the pentode is a vacuum tube which has a high value of plate resistance, the Norton's equivalent circuit must be used. Figure 54-25A shows the basic equivalent circuit. It shows all of the values of capacitance and resistance present in the circuit. In section 20-21 it was shown that Figure 54-25B is the low frequency equivalent circuit with the output taken across $R_g$. To the low frequencies, the output capacitance, $C_o$, the wiring capacitance, $C_d$, and the input capacitance, $C_{in}$, are negligible values; and are omitted from the low frequency equivalent circuit. The components which affect the low frequency response are the value of the coupling capacitor, $C_c$, and the value of the grid resistance, $R_g$. The equivalent resistance, $R_{eq}$, is the parallel resistance of the low value of plate load resistance, $R_L$, and the high value of plate resistance, $r_p$. This value, $R_{eq}$, is very small compared to the value of the grid resistance, $R_g$. The cathode resistance, $R_k$, is not shown because of its small value. Therefore, the low frequency response of the pentode RC coupled amplifier shown in Figure 54-24B is controlled by the value of the coupling capacitor and the value of the grid resistor.

At the high frequencies, the output capacitance, $C_o$, the wiring capacitance, $C_d$, and the input capacitance, $C_{in}$, become significant shunting values. They are lumped as the total capacitance, $C_t$, in Figure 54-25C. The coupling capacitor, $C_c$, appears as a short to this high

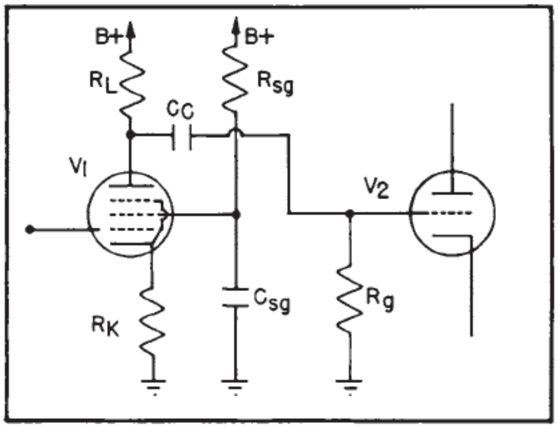


Figure 54-24 - Typical RC coupled pentode amplifier.

frequency. The resistance, $R_t$ is the combination of the shunt resistance of $r_p$, $R_L$, and $R_g$. Since $R_L$ is by far the smallest value (conventional in a pentode amplifier) the total shunting resistance, $R_t$, will also be a small value. Therefore, the high frequency response of the RC coupled pentode amplifier is controlled by the values of the shunting capacitance, $C_t$, and the value of the plate load resistance, $R_L$. Figure 54-26 shows the response curve for the pentode amplifier in Figure 54-24. Note how the response "falls off" at the high and low frequencies. Therefore, the circuit will distort the square wave output from an amplifier of this type. One of the ways of extending the response of the amplifier is to, in some way, increase the value of the gain bandwidth product for the circuit.

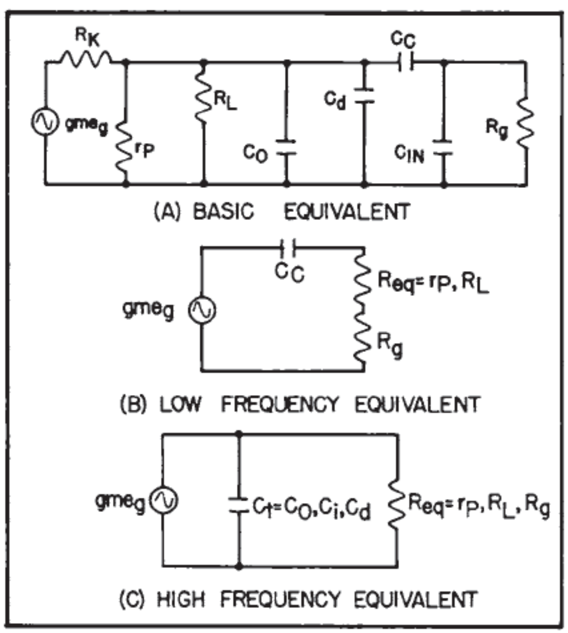


Figure 54-25 - Equivalent circuits.

Q15. What controls the low frequency response of a conventional RC coupled pentode amplifier used as a video amplifier?

Q16. With reference to $C_{in}$, what advantage do pentodes have over triodes at high frequencies?

## 54-14. Compensation

The frequency response of the ordinary amplifier shown in Figure 54-24 can be extended by use of the appropriate compensating networks. The high frequency response of the amplifier may be extended by the use of either SHUNT COMPENSATION, SERIES COMPENSATION, or a combination of both types called COMBINATION COMPENSATION.

There are various methods of extending the range of a video amplifier at the high frequency end of the range, but perhaps the simplest and most effective is the shunt-peaked method, shown in Figure 54-27A. As mentioned in a previous section, the gain at high frequencies is reduced because the load is shunted by $C_o$, $C_d$, and $C_{in}$. These same values can be made to extend the range if a small inductor, $L_1$, is inserted in series with the load resistor, $R_L$, to form a parallel resonant circuit. If the value of $L_1$ is properly chosen so that the circuit will be in resonance at the point where the response curve begins to fall appreciably, the range can be extended. The value of $L_1$ is critical. If the value is not correct, the amplification may be increased before the point at which the response curve begins to fall, which results in frequency distortion.

Series compénsation may also be used to extend the range of the high-frequency end, as indicated in Figure 54-27B. In this application,

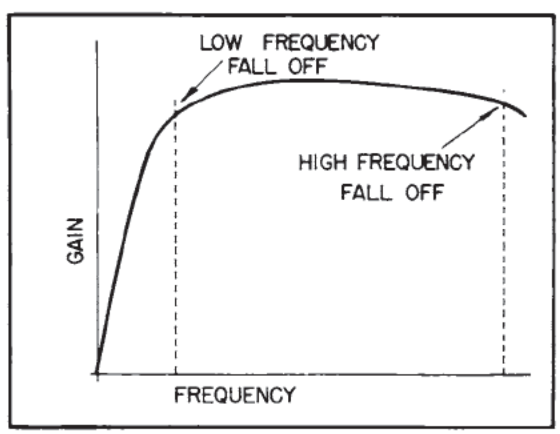


Figure 54-26 - Response curve.

A15. Primarily the value of the coupling capacitor and grid resistor.

A16. Because of reduced Miller effect, the $C_{in}$ of the pentode would be smaller and high frequency response would be improved.

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an inductor, $L_2$, of the proper value is added in series with the coupling capacitor, $C_c$, so that the series resonant circuit is formed with the parallel combination of $C_d$ and $C_{in}$. At resonance, increased current will flow through these capacitances and larger output voltages will be applied between the grid and cathode of $V_2$.

The high-frequency peaking effect of shunt compensation, in addition to the increased gain of series compensation may be obtained if both of these methods are used in the same coupling circuit. There are other factors, however, such as the transient response, which have to be considered in a network such as this.

At low frequencies, the distributed, or wiring capacitance may be neglected, but the reactance of the coupling capacitor becomes increasingly important. Since the reactance of this capacitor is:

$$X_C = \frac{1}{2\pi fC} \qquad (10\text{-}24)$$

it becomes appreciable at low frequencies. Consider a voltage divider made up of $C_c$ and $R_g$ as shown in Figure 54-28 in which the reactance of $C_c$ is large, as it would be at low frequencies. More of the voltage would appear across $C_c$ and less would appear across $R_g$. Since in some practical applications the voltage gain must be maintained within 70% of the mid-frequency gain, this loss at low frequencies can not be tolerated.

The capacitance of $C_c$ could be increased, but such a procedure would increase the stray capacitance and thus cut down the high-frequency gain.

Another factor that is perhaps more important than the loss of gain is the large phase shift which occurs at these frequencies. If 10 stages of video amplification are used, a phase shift (lead) of about $2^o$ is all that can be permitted per stage.

The phase shift can be reduced by employing a large value of coupling capacitance and the largest possible grid leak resistance; but both of these expedients have their disadvantages also.

Amplifiers that do not have as critical requirements as video amplifiers may be made to operate satisfactorily by using large cathode and screen bypass capacitors and a coupling capacitor as large as possible. Video amplifiers, however, require special compensation at the low frequencies.

Both the loss in gain and the increase in phase shift may be corrected by dividing the load resistance into two parts and bypassing one part with a capacitor. A circuit employing this method of low frequency compensation is shown in Figure 54-28.

The load resistance is made up of two parts, $R_L$ and $R_c$, of which $R_c$ is bypassed by C. At the higher frequencies, the load is effectively $R_L$ because $R_c$ is bypassed by C which has a low reactance at these frequencies. At low frequencies, however, C offers a high impedance and the load is effectively $R_L$ plus $R_c$. This effective load increase causes a greater proportion of the plate signal to appear as output and thus counteracts the normal drop at the low

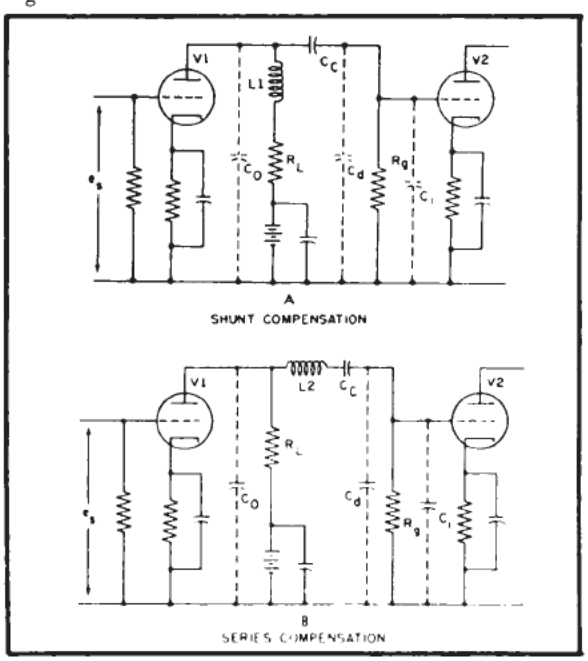


Figure 54-27 - High-frequency compensation.

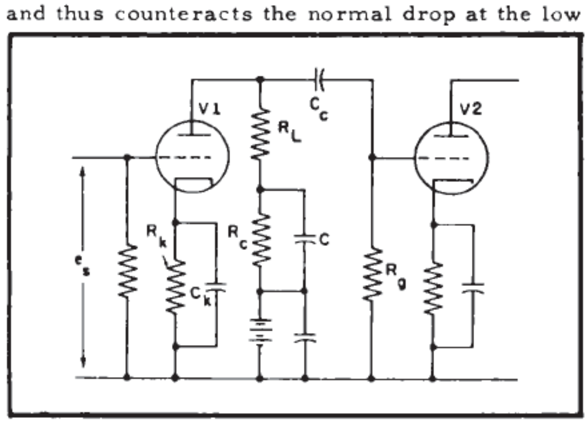


Figure 54-28 - Low-frequency compensation.

frequencies. Care must be taken in selecting the values of $R_C$ and C so that uniform gain can be extended into the lower frequencies beyond the point where the gain begins to fall off. Distortion will occur if the increase in gain takes place before the curve begins to fall off.

Another components influencing low-frequency gain is the cathode bypass capacitor. In order to prevent degeneration from occurring at the lower frequencies, becuase of inadequate shunting, the capacitance of the cathode bypass capacitor must be great enough to offer low impedance (with respect to the cathode resistor) at the lowest frequency to be amplified. Thus in video amplifiers, the capacitance of cathode bypass capacitors are many times greater than those used for audio amplifiers.

Q17. Why is the inductor in series with $R_L$ referred to as the shunt peaking coil?

## 54-15. Output Circuits

After the video pulse has been amplified by a compensated video amplifier, it is almost ready to be sent to the presentation system where it will appear as target information on the indicator associated with each receiver. Oftentimes, the presentation system is located at quite a distance from the output of a video amplifier. It is for this reason output circuits are used. If the presentation system is located several feet from the output of the video amplifier, it must be connected with coaxial cable. To do this with a maximum transfer of power, an amplifier circuit called a CATHODE FOLLOWER is used. The circuit is shown in Figure 54-29. A detailed analysis of this circuit can be found in Chapter 44. Briefly, the distinguishing feature of a cathode follower is the output is taken from the cathode circuit. The output voltage will be less than input voltage. The cathode follower can be used for isolation, impedance matching and power amplification.

The output impedance of the video amplifier will be high in comparison to the impedance of coaxial cable. The cathode follower can be designed to match these impedances and maintain the frequency response established in the preceding circuits.

Q18. Why are cathode followers not used as voltage amplifiers?

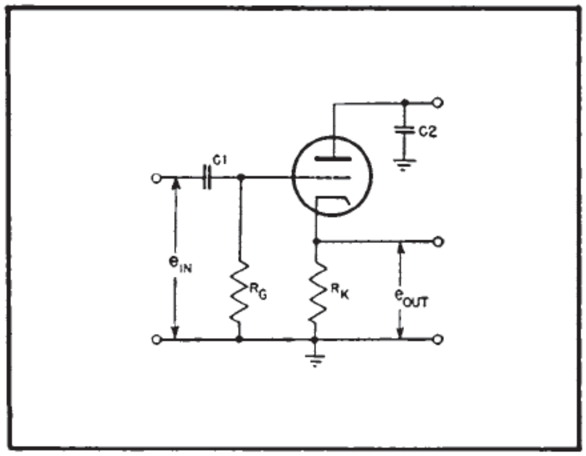


Figure 54-29 - Cathode follower.

A17. The shunt peaking coil is in parallel with the shunt capacitance and forms a parallel resonant circuit with this capacitance.

A18. Because the ac voltage at the cathode will be less than the ac voltage on the grid.

## EXERCISE 54

1. Describe the basic block diagram of the radar receiver?
2. Draw the output waveform of the IF strip of a radar receiver.
3. Why must a non-linear device be used for mixing action?
4. Describe heterodyning.
5. What is a local oscillator? What type of output does it provide?
6. What is detection, as applied to a radar receiver?
7. Describe the requirements for a radar receiver local oscillator.
8. Compare velocity modulation to density modulation.
9. Describe the operation of the basic velocity modulated tube.
10. What is the purpose of cavity resonators?
11. What is the function of buncher grids?
12. How does the basic klystron differ from the basic velocity modulated tube?
13. What is the reason for the half wavelength space between the catcher grids?
14. How does the reflex klystron differ from the basic klystron?
15. When does bunching occur in the reflex klystron?
16. What is the function of the repeller plate?
17. What controls the frequency of the reflex klystron?
18. Where would AFC be applied to the reflex klystron?
19. What is meant by the term "mode of operation?" How are the different modes of operation of the reflex klystron determined?
20. What type of device is used in a radar mixer?
21. Describe the operation of the silicon diode.
22. How does the balanced crystal mixer differ from the single ended crystal mixer?
23. How does the balanced crystal mixer operate?
24. What is the difference between the use of a waveguide and a coaxial line when they are used for balanced crystal mixers?
25. Why is noise canceled in the balanced crystal mixer?
26. What is the purpose of the IF strip?
27. Why is a cascode amplifier used as the IF input stage?
28. What is the gain-bandwidth product? Of what use is it?
29. What is transient distortion? How can it be prevented?
30. What are swamping resistors? What are they used for?
31. What is the purpose of detection in a radar receiver?
32. Why is a conventional pentode RC coupled amplifier unsuited for use as a good video amplifier?
33. What is meant by high frequency and low frequency compensation?
34. Describe the methods of frequency compensation and why they work.
35. When is a cathode follower used in the output of a radar receiver?

# CHAPTER 55

# RADAR INDICATORS

The purpose of the radar indicator is to display the received signals from the radar receiver by producing a visual indication of returning echo information. The cathode ray tube is an ideal instrument for the presentation of radar data since it not only shows a variation of a single quantity such as voltage, but gives an indication of the relative values of two or more synchronized variations. The usual indicator is basically the same in function as the low-frequency test oscilloscope. The focusing, intensity and positioning controls are similar. The sweep frequency of the radar indicator is determined by the pulse repetition frequency of the system. The sweep time on the indicator's cathode ray tube is established by the settings of the range selector switch.

## 55-1. Basic Functional Block Diagram

Synchronization of events is the most important aspect of the radar receiver-transmitter. It is particularly important in the radar presentation system. At the instant of time that the transmitter fires, the circuits which control the presentation on the indicator must start. This instant or time zero is usually noted on the indicator by a bright spot at the start of the sweep or a positive pulse depending on type of indicator.

To insure accurate range determination, these events must be performed with the highest degree of accuracy.

The basic block diagram for a presentation system is shown in Figure 55-1. It is composed of a sweep circuit, a GATE CIRCUIT, and a cathode-ray tube (CRT). The inputs to the circuit are as shown in the diagram. From the radar timer, a synchronization pulse is sent to both the sweep circuit and to the gating circuit. It will cause a sawtooth of voltage to be developed by the sweep generator, and a square wave pulse output from the gating circuit. Both of these waveforms are then applied to the CRT. It should be noted at this time that the duration of the sweep across the scope will be determined by the duration of the gating pulse. Video information from the receiver will be sent directly to the CRT where it will be presented on the scope.

Q1. What type of waveshape is applied to the CRT from the receiver?

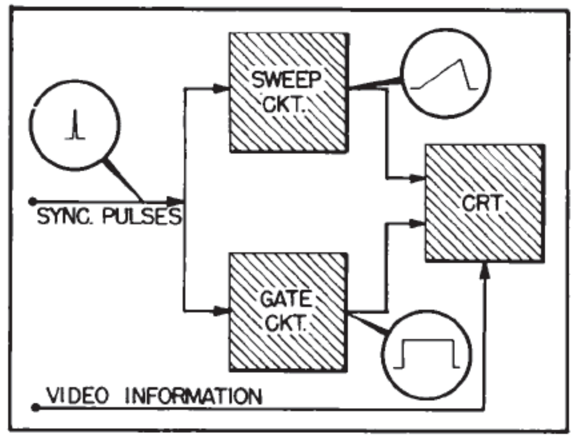


Figure 55-1 - Basic block diagram of presentation system.

## 55-2. Basic Principles of Operation

The function of the sweep circuit is to produce a sweep voltage which will subsequently be applied to the CRT. This sweep circuit will operate in much the same way as those described in Chapter 43.

In order to prevent the appearance of the retrace pattern on the screen of the scope, a gating pulse, supplied by the gate circuit, is applied to the CRT. This gating pulse permits electrons to flow in the cathode-ray tube only during the time of the forward sweep of the sawtooth wave. No electrons are permitted to strike the screen during the flyback, or return-trace time, because this pulse, in a sense, opens and closes a gate through which electrons must pass, it is sometimes called an INTENSITY GATE voltage. The type of circuit used for the gate circuit may be a multivibrator.

Q2. What is the purpose of the gating pulse?

## 55-3. Electromagnetic Deflection

In most modern presentation systems, the electromagnetic deflection of the cathode-ray tube electron beam is preferred in lieu of electrostatic deflection. The reasons for this choice are: increased control of the electron beam, improved deflection sensitivity, better

A1. A video pulse.

A2. To make sure that the tube sweeps only when the gating pulse is coincident with the sweep signal. This prevents a visible retrace on the CRT tube screen.

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electron beam positioning accuracy, and simpler construction. The primary difference between the electromagnetic and the electrostatic types is the method used for deflection and focusing of the electron beam. Both types employ an electron gun, and use an electric field to accelerate and control the flow of the electron beam. Physically, the cathode-ray tube employing electromagnetic deflection is constructed much like the electrostatic type. This is shown in Figure 55-2.

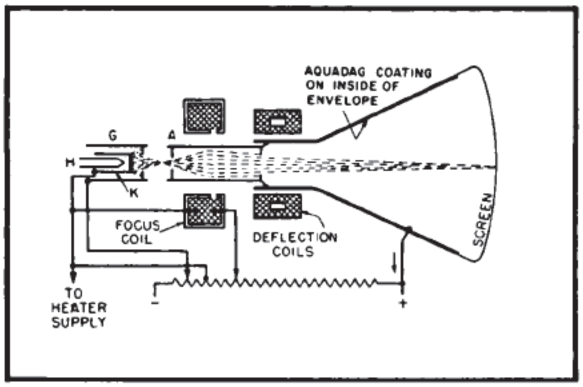


Figure 55-2 - Basic electromagnetic CRT.

The type of electron gun used in the electromagnetic CRT is the triode gun. It is composed of the normal heating element, a cathode and a control grid. Beyond the control grid is the accelerating anode which contains several baffles. It is a hollow cylinder which is coaxially symmetrical with the control grid cylinder. The accelerating anode is connected to the conducting coating of the tube. This coating acts as an extension of the accelerating anode and as a shield. Beyond the accelerating anode, there are two elements called FOCUSING COILS. With the magnetic fields produced by these coils, the focusing of the electron beam will be affected.

If the tube is operated, a small glowing spot will be seen on the face of the tube. If focusing and deflection adjustments are properly made, the spot will be a small speck in the center of the screen. It will not move until either the focusing or deflection fields are changed. One other important component part of CRT's employing electromagnetic deflection should be mentioned here. When the tube is operating, electrons are released from the cathode at a high velocity. These high velocity electrons when striking tube elements such as the control grid, or when striking the atoms of the gas included within the tube, can cause negative ions to be produced. Since the mass of the negative ion is considerably more than that of the electron, the deflection and focusing fields will have little or no control over their movement. Therefore, the negative ions will move toward the screen and strike it causing a glowing spot. Since the fields have little control over its position, the spot on the face of the scope will remain fixed in one position. Although a small spot is produced, over a long period of time, a hole will be burned in the phosphorous coating. To reduce these effects devices called ION TRAPS are used. In the electromagnetic type of tube the electron stream, including the negative ions, is purposely directed toward the side of the tube. An ion trap magnet is placed around the neck of the tube to deflect the electrons back to the axis of the tube. The path of the heavier negative ions will be relatively undisturbed by the field from the ion trap magnet, and they will move toward, and be collected by, the highly positive anode. The ion trap magnet is normally located between the control grid and the accelerating anode. Its position is critical, and any adjustment performed on it must be done carefully.

Focusing of the electron beam on the face of the screen is accomplished by use of the focusing coils. The focusing coil acts in much the same way as a lens. Actually, the focusing is caused by the use of two lenses. The first of these lenses is produced by the electrostatic field between the control grid and the electrode following it. This action is similar to the focusing of the electrostatic type of deflection. A cross over point or point of divergence is caused to exist in the region near the focusing coils. The fields about the focusing coils is uniform. Since the field is uniform, there is a force exerted on the electrons by the magnetic field which causes them to move at right angles to their direction of motion. Since the electron remains within the field and feels a constant force at every point, the path of the electron tends to be circular. If the strength of the magnetic field is made stronger, the circle becomes smaller. If the field is weaker, the circle diameter increases. Since the electron possesses forward motion, the pattern described by the electron in moving through the uniform field will be helical as shown in Figure 55-4. The helical pattern will be projected by all electrons entering the field at some angle with respect to the direction of the magnetic field. If an electron enters the field and moves parallel to the line of

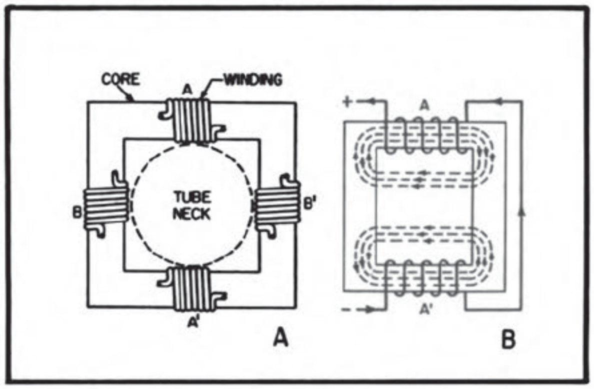


Figure 55-3 - Yoke schematic

force it will not be deflected by the field. It can be seen then that by choosing the proper strength of the magnetic field about the focusing coil, the electrons entering the field at an angle and those entering parallel to the field can be caused to arrive at the face of the screen in a very narrow beam. For proper operation of the tube, the magnetic field acts on the diverging electron near it in such a manner that they describe a single turn helical path with the result that the new cross-over point for electrons leaving the vicinity of the focus coils is at the face of the scope. To control the magnetic field which in turn controls the cross-over point of the electrons, the current through the focus coils must also be controlled. This is usually done by the use of a rheostat called the FOCUS CONTROL. With this control the small spot can be produced on the screen. The beam may now be deflected.

To discuss electromagnetic deflection, the beam after passing through the focus coils will be assumed to be very narrow. Any action of the magnetic field upon the electron beam will be described assuming that the beam is a single electron.

The magnetic fields which deflect the beam are produced by causing currents to flow through DEFLECTION COILS mounted around the outside surface of the tube. Normally, there are four deflection coils used. Two of these, wired in series, are positioned in a manner to cause the magnetic lines of force produced by them to be in a vertical direction. The two other coils, also connected in series, are positioned so that the magnetic fields produced by them are in a horizontal direction. This means that the coils combined will produce two fields which are effectively 90° out of phase or perpendicular to each other. The coils which produce the field which is horizontal are called the VERTICAL DEFLECTION COILS. The other two are naturally called the HORIZONTAL DEFLECTION COILS because they will produce fields which will cause the electron to move across the face of the screen.

The coils are mounted about the outside surface of the tube in one complete assembly called a YOKE. It is located beyond the focus coil, and immediately before the flanged section of the tube enclosure. Schematically, the coils may be pictured as shown in Figure 55-4.

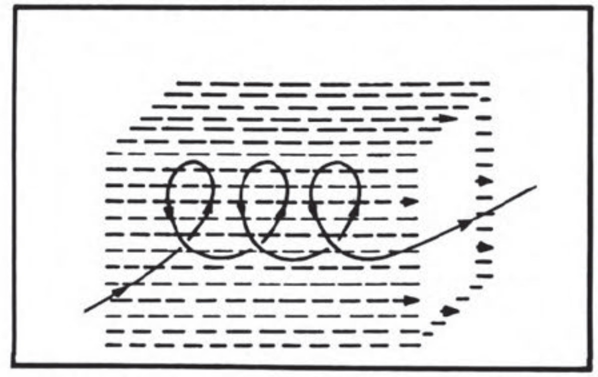

Figure 55-4 - Helical motion of electron in passing through uniform magnetic field.

Electron deflection in the electromagnetic CRT is accomplished in the same manner as it was in the electrostatic type with the exception that the deflection in the magnetic type is caused by the density of the field. If the current through the horizontal deflection coils is increased, the field produced by both of the horizontal coils will increase resulting in a greater horizontal deflection. The same increase in the current flow through the vertical deflection coils will produce the same amount of vertical deflection.

Obviously, the type of current that would cause a sweep to exist across the face of the scope could not be a constant dc current, it must be one that increases linearly, because the electron must be moved across the screen at a constant velocity to produce the visible base line that is known as the sweep. One type of current waveform that would accomplish this function is some type of sawtooth waveform. A sawtooth is a constantly increasing linear current or voltage. If a sawtooth is then applied to the horizontal deflection coils, the electron would be deflected. However, the resulting deflection would not cause a linear line to be traced across the face of the scope. The type of waveform that would accomplish this effect is known as a trapezoidal waveform. The trapezoidal waveform and its generation will be considered in a subsequent chapter. It is sufficient to know at this time that the type of wave applied to the horizontal deflection coils resembles a sawtooth current waveform. It is also important at this

time to know that it is the change in magnetic field density which is responsible for the deflection of the electron, or the electron beam.

Q3. What property of electrons allow them to be deflected by a magnetic field at right angles to their motion?

55-4. Azimuth Sweep Deflection

In many search type radar systems, two or more types of scan are used. One of the more popular combinations is the "A" scan and the "PPI" scan. The "PPI" is used to locate the approximate position of the target while the antenna is scanning through 360°. One of the other more popular displays is a "B" scope. With the "B" scope, a target's position to the right or to the left of a predetermined bearing can be determined. Its actual range can then be found directly on the "B" scope. Which ever of these combinations are used a problem relative to both must be resolved. The problem is to synchronize the rotation of the "PPI" scan or the "B" scan with the rotation of the antenna. This can be accomplished through the use of devices called SERVO MECHANISMS. These devices will let the scope scan to know exactly where the antenna is scanning, or will control the scan of the scope. This process can be performed either electrically or mechanically. The yoke of the electromagnetic CRT can be mechanically connected to the antenna so that it will rotate about the neck of the tube at exactly the same speed as the antenna. This system has mechanical disadvantages. The electrical method used to synchronize this rotation is affected by applying signal of the proper phase displacement to the horizontal and vertical deflection coils. The synchronization of these types of scan will be explained thoroughly when information concerning the operation of servo mechanisms is understood.

Q4. What sort of information can be taken from the "PPI" scope? From the "B" scope"

## EXERCISE 55

1. Why is synchronization important in a radar presentation system?
2. What is a "main bang"?
3. What is the function of a gate circuit?
4. Describe the visual symptoms of an inoperative gate circuit.
5. Describe the function and operation of the focus coils in an electromagnetic CRT.
6. Why does an electron describe a helical path when passing through a uniform magnetic field?
7. Describe how deflection is accomplished in the electromagnetic CRT.
8. What type of current is necessary to cause horizontal trace on the CRT? To what is this current applied? What would happen if this current would cease?
9. What is a "PPI"?
10. Describe how synchronization of antenna movement with "PPI" scan is affected?

A3. The property of deflection possessed by this system is due to the interaction of two fields--the field about the deflection coils and the field about each individual electron.

A4. The "PPI" scope will give range and azimuth information for 360°. The "B" scope will give range and direction within 90° of dead ahead.

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